ることができる。

第３５図は、各ビームで利用できる黒、赤および緑色で示す（周波数、タイムスロット、コードもしくはそれらの組合せとすることができる）３チャネルもしくはチャネル群に対する単純化された例を示す。（例えば、－４．５ｄＢの）設計交差点におけるビームエッジは第３５図に大きなカラー円で示されている。

したがって、黒の大きな接触円は、“黒”チャネルを使用するビームを表しピーク利得から－４．５ｄＢ下で接触している。大きな赤の接触円は、赤チャネルに対するビームパターンを表す。これらは“黒”ビームに対して変位されており、この固定変位は例えば“赤”チャネルに対するフェーズドアレイの位相を“黒”チャネルに対して修正することにより達成される。“黒”チャネルに非修正ビームパターンが使用され、３つの隣接給電の各々に“赤”セル用エネルギの１／３が送られ、マルチ給電パラボラ反射器によりそれを達成することもできる。コヒーレント付加により、“赤”セルの中心における利得はその点における“黒”ビーム利得の３倍となり、有効に“孔が埋められる”。“緑”周波数すなわちタイムスロットに対して“緑”ビームが全く同様に形成される。これは各給電に直接関連するトランスポンダチャネルを介して適切な信号の組合せを指令する地上ハードウェアを使用して達成される。

第３５図において、小円はそのそこまでは特定のチャネルが使用されるエリアを示し、それを越えるとより中心指向された別のビームを利用することができる。

このエリアは黒ビームの場合には埋められて識別し易くされる。このエリアはビーム中心から外へビーム半径のルート３を越えて伸びている、すなわち“セル”エリアは“ビーム”エリだアの僅か１／３に過ぎず、“セル”内の移動体はビーム半径の５０％を僅かに越えるだけの“ビーム”しか使用しない。

もちろん、実際には、ビーム当たり３チャネルよりも遥かに多くのチャネルを使用することができ、したがって利用できるチャネル数をＭとした場合ビームスポット面積の１／Ｍしかないセルを計画することができる。例えば、Ｍ＝７であれば、第３６図に示すように、ビームはその半径の１／ルート７までしか使用さ

れない。実際にはMは少なくとも１００であり、セル半径はビーム半径の１／１０となることがあり、したがってビーム構成の利得およびC／I性能が重要となるのはビームスポット覆域の小部分にすぎない。“孔を埋る”ことが容易ではなくなるため、これは必ずしもスポットを縮小して利得を高くできることを意味しない。多数のオフセットビームにより、第２４図および第３０図に示すように、物理的給電の位相をピーク利得がどこでも得られるようにするのが望ましい。（２つの給電だけの位相調整により）利得を得ることが最も困難な場所は２つのスポットの中間であり、このような状況における利得は第２４図および第３０図に示すようにビームエッジ交差点を選択することにより最大とされる。２つのビーム間の利得は、（例えば、３ｄＢ以上の）ビームエッジ利得の２倍となり３つのビーム間の利得はその点における一つのビームの３倍となる。これはこれら３つの点において利得を比較するために第２４図および第３０図で使用される校正を説明するものである。

したがって、いずれの場合にも、電力効率の理由から熟慮される１／２～１／３のレートのエラー修正コーディングにより、前記した再利用分割技術を利用する場合、各ビームにおける即時周波数再利用により得られるＣ／Ｉの許容範囲を得ることができる。再利用分割技術によりゼロ生成や干渉キャンセルを頻繁に行うことなくこれが達成される、すなわちアンテナの自由度の全てが利得を最大とするのに使用される。干渉キャンセルや近隣セルの中心にゼロパターンを生成する技術をさらにボーナスとして利用して、近隣ビームからのＣ／Ｉを無視できる割合まで低減することができる。

本発明のこの実施例を実現するのに使用できる代表的なコーディング方式は、１／４もしくは１／５レートに基づいた破壊（punctured）たたみ込みコーディングであり、各未符号化音声のコーディングはその知覚的意味に従って、例えば、１／２～１／５レート間のレベルへ適応される。ＢＰＳＫではＱＰＳＫよりも３ｄＢ低いＣ／Ｉが許容されるが、２：１帯域幅効率損失を被る理由は見当たらない。２倍のコーディングにおけるＱＰＳＫのＣ／Ｉ許容範囲は事実半分のコーディングのＢＰＳＫよりも良く、したがって少なくとも下りに４相変調を使用す

ることができる。

前記した検討は、極めて広帯域のＴＤＭを使用する衛星から移動体へのチャネルでは達成できるが狭帯域ＦＤＭでは達成できない、コヒーレント復調性能に基づいている。下りの方法に対する基準は、１００Km／Hrの車速で２．５GHzにおいておよそ２００μSである、チャネルのフェージング成分がスタティックであると見なされる時間中復調および復号すべき情報ビット数が多くなければならないということである。したがって、情報レートは５Kb／ｓよりも数桁高くなければならず、例えば、１／３の平均コーディングレートの場合送信ビットレートは、４相変調を使用する１MHzの帯域幅チャネルを通過する、１．５Mb／ｓよりも高くする必要がある。前記した技術に基づくシステムにより提供される容量は、４Kb／ｓベクトルを使用する場合はMHz当たりスポットエリア当たり１００アーラン程度であり、２．４Kb／ｓを使用する場合はMHz当たりスポット当たり１６６アーランである。

本発明の前記した再利用方式を実現する一つの代表的な技術は、前記した、地上ビーム形成によるものである。これには信号間の相対位相および振幅差が持続されるように各アンテナ給電へ信号を運ぶフィーダリンクを地上局から衛星まで設けることが含まれる。このようなコヒーレントトランスポンダを使用すると、アンテナの給電点当たり衛星に必要なトランスポンダチャネルは一つだけである。

固定ビームの場合に本発明を実現する別の手段が以下に記載され、衛星にさらにハードウェアの費用をかけるコヒーレントフィーダリンクの必要性が回避される。固定していて変化しなければオンボードビーム形成を使用するのは最も簡単であり、固定エリアを照射する静止衛星に良く適している。非静止衛星についても本発明のこの実施例を利用することができるが、その場合にはビームが固定エリアを照射するように衛星の動きを補償する計画的に適応されたビーム形成の利点を得ることは容易ではない。

利用できる全帯域幅がＮのサブバンドへ分割されその各々が第３５図に示すようなセルラー再利用パターンに従って地上のエリアを照射するのに使用される、

ＦＤＭＡに対する固定ビーム形成トランスポンダの実現を第３７図に示す。第３５図に黒、赤および緑で示す３つのサブバンドを使用して説明を行う。

１組のトランスポンダチャネル３７により対応する１組のフィーダリンクから信号が受信され適切な中間周波数へダウン変換されて増幅かつ濾波される。３７１０のＩ．Ｆ．出力はＩ．Ｆ．信号の重みづけされた組合せを形成するＩ．Ｆ．ビーム形成網３７２０へ加えられる。“黒”チャネルは未修正アンテナパターンに直接対応するように任意に選択される、すなわち、黒信号１はアンテナ給電１番だけから直接放射しなければならず、黒信号２はアンテナ給電２番だけから放射しなければならず、以下同様である。したがって、ビーム形成網により単位重み付けの黒チャネルは指示されたアンテナ給電に対応する加算網のみに接続される。

しかしながら、赤チャネルおよび緑チャネルは３つの黒ビームの中間に中心を有するビームパターンで放射しなければならない。したがって、黒ビーム１，２および３の中間にある赤ビームは１／ルート３の電圧／電流重み付け（１／３の電力重み付け）を介して関連する３つの加算網に接続される。したがって、“赤”エネルギの１／３は所望の“赤”中心を取り巻く３つの給電の各々を介して放射される。同様に、黒ビーム２，３および４の中間にある緑ビームは１／ルート３の重み付けを介して給電２，３および４に関連する加算器に接続される。前記した重み付けは代表例であり説明の目的で単純化されている。Ｉ．Ｆ．ビーム形成網は本質的に主に単純な抵抗素子からなる網により実現できるため、複雑さの影響を許容できる範囲でより複雑な重みセットを使用することができる。例えば、４つ以上の隣接給電に給電を行ってビームを形成することができ、かつ負の重みを使用して所望の場所における放射パターンにゼロを生成することができあるいはＣ／Ｉを増大するためにサイドローブレベルを低減することができる。

抵抗性Ｉ．Ｆ．ビーム形成網を形成する一つの方法は、絶縁基板上に堆積された連続シートすなわち薄膜状の抵抗材料を使用することである。概念的にこのシートはビームにより照射される２次元表面に対応すると見なされる。第３５図に示すように、“黒”ビーム信号に対応する信号電流が“黒”セルの中心に対応し

て配置された点でシートへ注入され、“赤”および“緑”信号電流が黒信号注入点とお互いの間の１組の点において注入される。第３８図に注入点をラベル’Ｉ’で示す。

黒、赤および緑の所望の組合せに対応する信号電流が黒、赤および緑の注入点の中間に配置された接点により抵抗面から抽出される。これらの電流抽出点は第３８図’Ｅ’として示されている。黒ビームに対する一つの重み１と赤および緑ビームに対する１／ルート３の３つの等しい重みを有する前例とは対照的に、この技術により黒、赤および緑ビームに対して同じ重み分布が与えられる。抽出された電流は“仮想アース”増幅器入力もしくはベース接地バイポーラトランジスタのような低入力インピーダンス増幅器へ送られる。この技術により実現される重みセットは電流注入および抽出コンタクトランドの形状およびサイズを選択することにより調整することができる。サイズおよび形状を決定するための単純なルールは提案されておらず、抵抗シートの流入電流および電位の２次元有限要素コンピュータ分析を実施して、提案を単純に確認しなければならない。

Ｉ．Ｆ．形成網から発生された組合せ信号は、アップコンバータ３７３０へ送られて所望の衛星ｰ移動体周波数帯域へ変換される。アップコンバータは信号の相対位相を持続するために全て同じ局部発信機信号により駆動され、かつ相対振幅を持続するために利得が整合されている。アップ変換された信号は行列電力増幅器３７４０により増幅して電力レベルを所望の送信電力まで高めることができる。

前記した本発明の技術を拡張して利用可能な総周波数帯域の再分割に関連する任意数の仮想ビームを発生することができる。３色の例では、各“色”は１／３のサブバンド幅に関連している。例えば、合計１６．５MHzを利用できる場合には、各トランスポンダチャネル帯域幅は公称５．５MHzとすることができる。例えば、給電数が３７であれば、３７の５．５MHz“黒”ビーム、３７の５．５MHz赤ビームおよび３７の５．５MHz緑ビームが発生される。したがって、通信に利用できる総帯域幅は、１６．５MH全部の即時周波数再利用を“黒”ビームだけに利用できる場合の、１６．５MHzの３７倍となる。したがって、本発明により

Ｃ／Ｉを著しく改善しながら即時周波数再利用パターンと同様に帯域幅を効率的に使用することができる。

このように利用できる特別な容量は、ＦＤＭＡバージョンの場合、トランスポンダチャネル数を増加することによって得られしたがってハードウェアも比例して複雑となる。次にハードウェアを複雑にすることなく容量が増大される代表的なＴＤＭＡ実施例を有利に構成する方法を示す。

第３９図に代表的なＴＤＭＡ実施例を示す。この場合トランスポンダチャネル３９１０の数はアンテナ給電数と同じであり、各チャネルの帯域幅はシステムに利用できる全帯域幅である。Ｉ．Ｆ．ビーム形成網３９２０も黒、赤および緑ビームの合成について前記したように機能するが、通信スイッチ３９１１によりトランスポンダチャネルセットには一時に１色しか接続されない。（１）全てのトランスポンダチャネルが対応する数の“黒”ビーム入力に接続されるか、もしくは（２）スイッチ３９１１を全て同時に作動させることにより、全トランスポンダが赤ビーム入力に接続されるかもしくは、第３９図に示すように、緑ビーム入力に接続される。

ＴＤＭＡフレーム期間の最初の部分（例えば、１／３）については黒ビームが使用され、第２の時間部分については赤ビームが励起され、第３の時間部分については緑ビームが励起されるようにスイッチが巡回される。スイッチが各位置にある期間は等しくなくてもよく、原則的に任意のセルにおいてどの色が最も高い瞬時容量を要求するかに従って適応させることができる。残りのトランスポンダの機能はＦＤＭＡについて前記したとおりである。

スイッチ３９１１の通信は中央地上局からの送信と同期化されているのが良く、それは、地上からプログラムして規則正しいスイッチ動作サイクルを実行し、マスタータイマーである衛星への地上局送信を同期化することができるオンボードクロックを設ける等の、さまざまな技術のいずれかにより達成することができる。また、地上局はトラフィックチャネルとは別の制御チャネルを使用してスイッチ

コマンドを送信することもできる。ビームの旋回と地上網との同期化を達成する

方法は本発明の原理にとって重要ではない。

前記した本発明のＴＤＭＡおよびＦＤＭＡバージョンでは共に固定ビーム形成網が使用されたが、本方法の明らかな拡張として所与の周波数やタイムスロットにより照射される地球のエリアを出来るだけ固定するように地上から制御されるスイッチ３９１１を使用してビームカラーへの周波数やタイムスロットの分配を置換することができる。もちろん使用する“色”（すなわち、タイムスロットやサブバンド）の数を増やせばこれは一層正確に達成される。サブバンドの数を増やすとＦＤＭＡの場合にはハードウェアが複雑となり、この点からはＴＤＭＡバージョンが好ましい。したがって、通信スイッチの同調は衛星の動きを補償して特定のタイムスロットや周波数により照射されるエリアが多少なりとも一定に保たれるように選択することができる。本発明は任意数のタイムスロットおよびサブバンドに応用することができ、後者の場合にはトランスポンダ信号のアナログ／デジタル変換、デジタル濾波およびデジタル重み乗算を使用するデジタルビーム形成を具備してデジタルに実現するのが有利である。

前記した実施例はあらゆる点で本発明を、制約するものではなく、説明するためのものである。したがって、当業者であれば本発明の細部のさまざまな変更を明細書から引き出すことができる。これらの変更や修正は全て請求の範囲に入るものとする。

【図１】

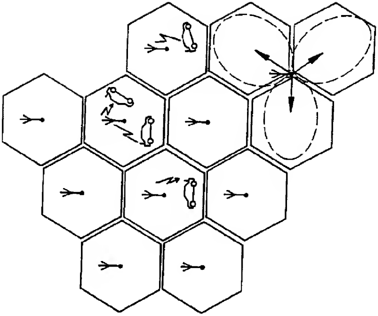


FIG. 1
(PRIOR ART)

【図２】

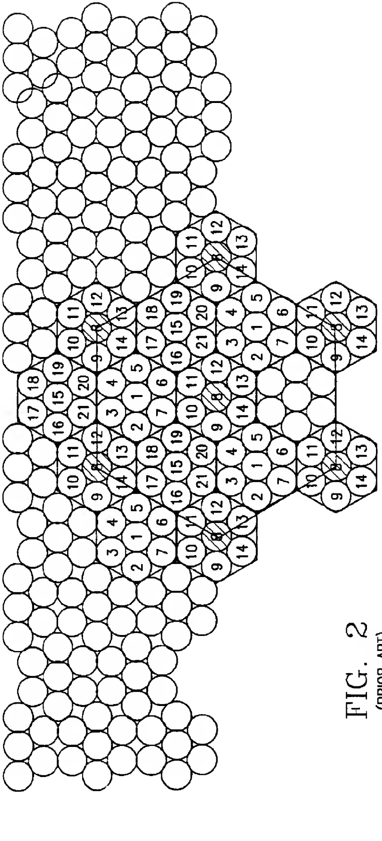


FIG. 2
(PRIOR ART)

【図3】

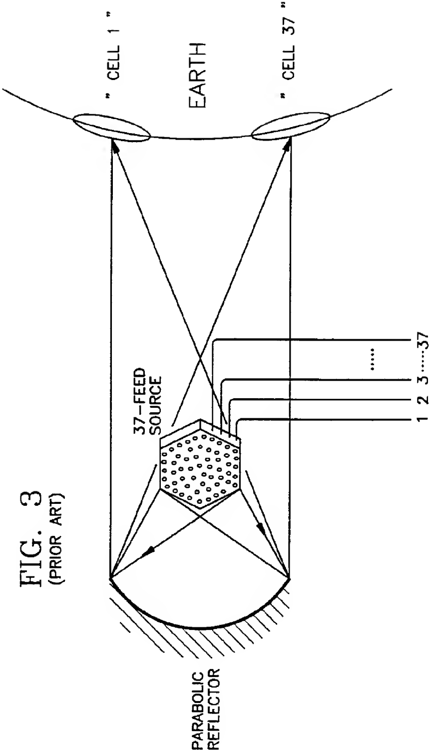


FIG. 3
(PRIOR ART)

【図４】

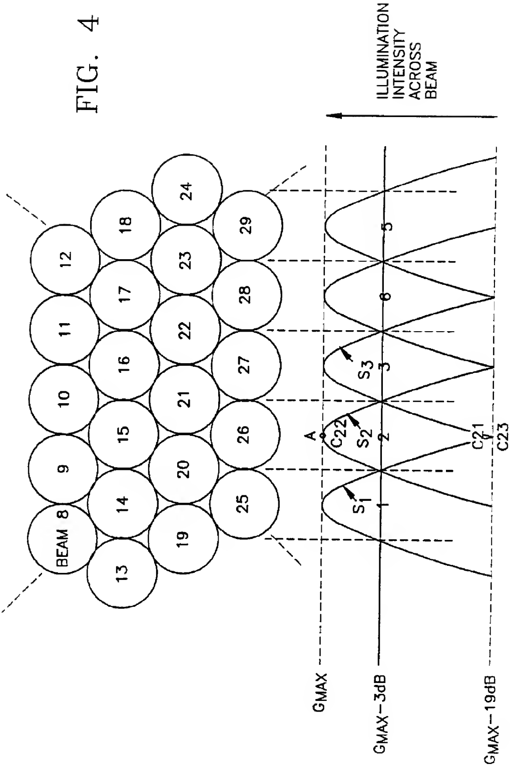

ILLUMINATION INTENSITY ACROSS BEAM
GMAX
GMAX−3dB
GMAX−19dB
BEAM 8
9
10
11
12
13
14
15
16
17
18
19
20
21
22
23
24
25
26
27
28
29
A
C22
C21
C23
S1
S2
S3
1
2
3
6
5

FIG. 4

【図5】

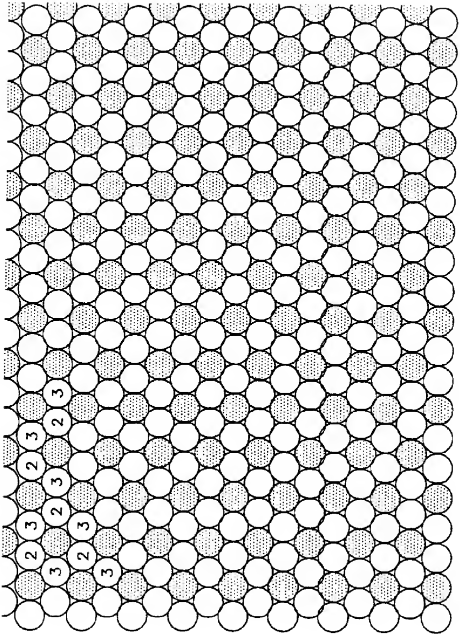


FIG. 5

【図６】

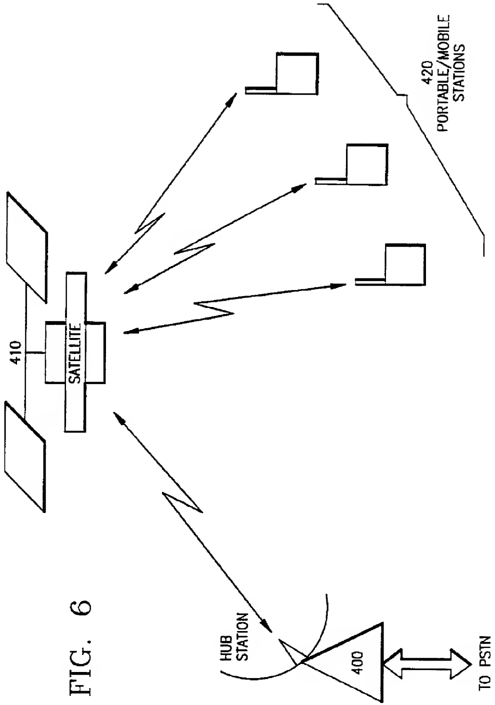
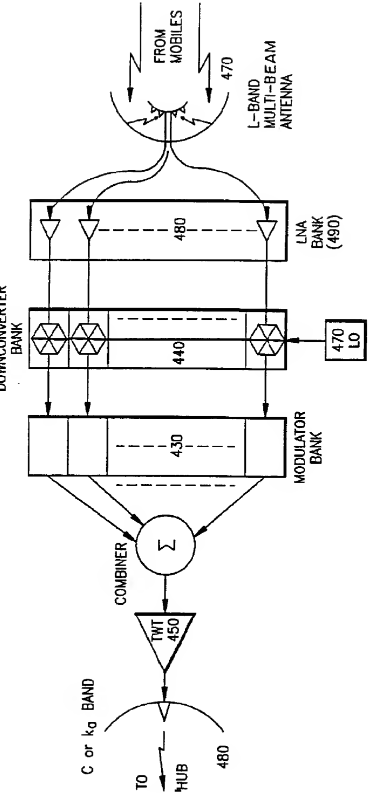

410
SATELLITE
420
PORTABLE/MOBILE
STATIONS
HUB
STATION
400
TO PSTN

FIG. 6

【図７】

FIG. 7


FROM MOBILES
470
L-BAND MULTI-BEAM ANTENNA
480
LNA BANK (490)
DOWNCONVERTER BANK
440
470 LO
430
MODULATOR BANK
COMBINER
Σ
TWT 450
C or ka BAND
TO HUB
480

【図8】

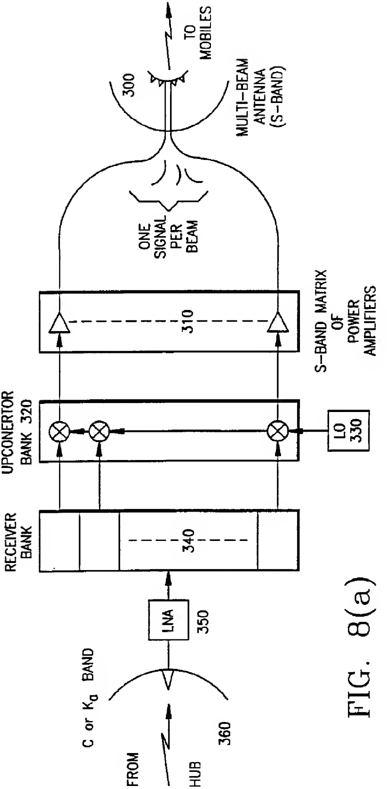

TO MOBILES
300
MULTI-BEAM ANTENNA (S-BAND)
ONE SIGNAL PER BEAM
310
S-BAND MATRIX OF POWER AMPLIFIERS
UPCONERTOR BANK 320
LO 330
RECEIVER BANK
340
LNA
350
C or K_a BAND
FROM HUB
360

FIG. 8(a)

【図８】

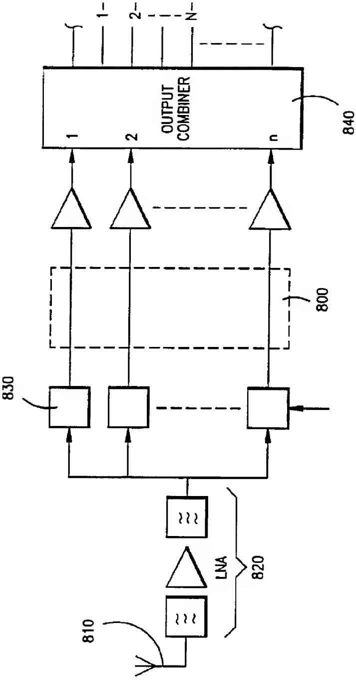


FIG. 8(b)

【図９】

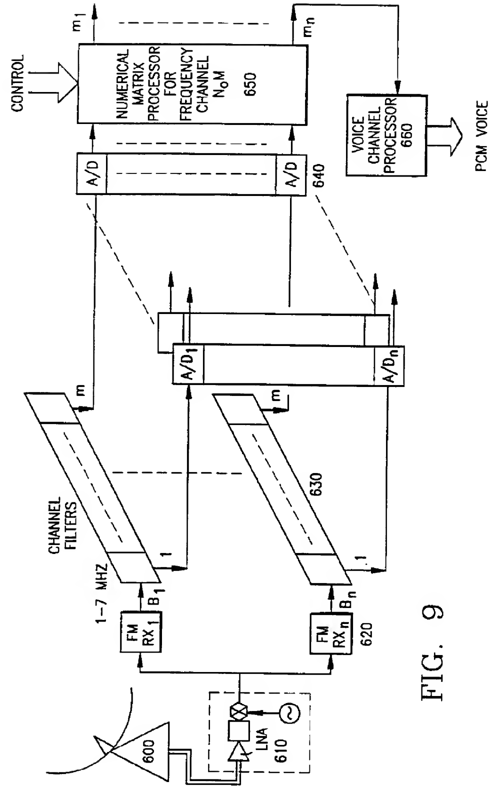

CONTROL
NUMERICAL MATRIX PROCESSOR FOR FREQUENCY CHANNEL N_oM
650
m_1
m_n
A/D
A/D
640
VOICE CHANNEL PROCESSOR
660
PCM VOICE
A/D_1
A/D_n
m
m
CHANNEL FILTERS
630
1
1
1-7 MHZ
B_1
B_n
FM RX_1
FM RX_n
620
LNA
610
600

FIG. 9

【図１０】

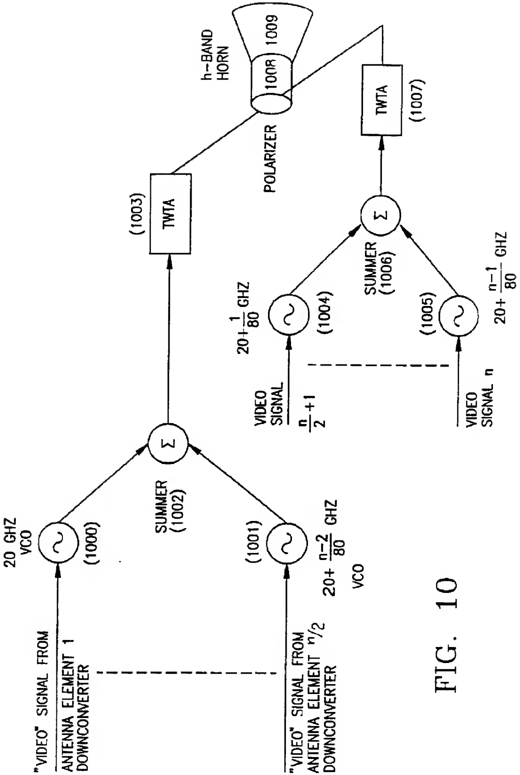

"VIDEO" SIGNAL FROM ANTENNA ELEMENT 1 DOWNCONVERTER
"VIDEO" SIGNAL FROM ANTENNA ELEMENT n/2 DOWNCONVERTER
20 GHZ VCO
(1000)
(1001)
$20+\frac{n-2}{80}$ GHZ
VCO
SUMMER (1002)
(1003)
TWTA
VIDEO SIGNAL $\frac{n}{2}+1$
VIDEO SIGNAL n
$20+\frac{1}{80}$ GHZ
(1004)
(1005)
$20+\frac{n-1}{80}$ GHZ
SUMMER (1006)
TWTA
(1007)
POLARIZER
h-BAND HORN
1008
1009

FIG. 10

【図１１】

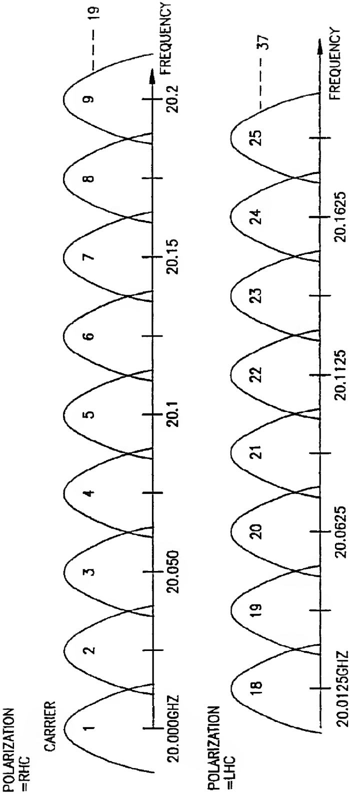
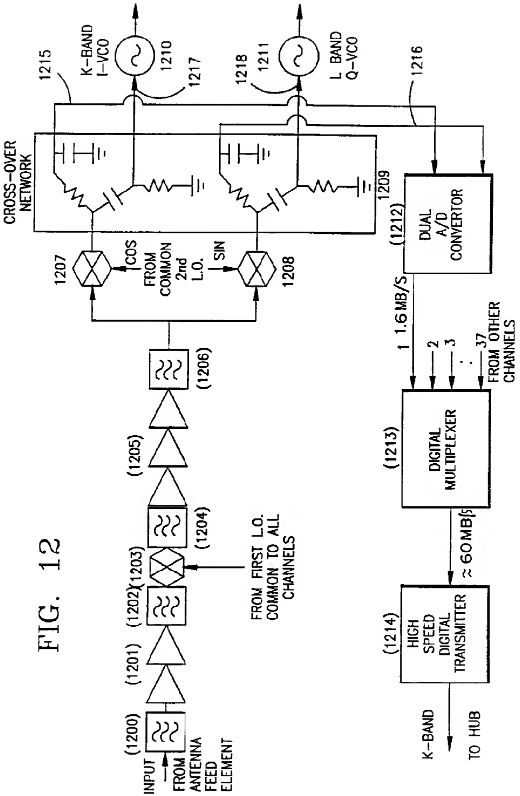

POLARIZATION =RHC
CARRIER
1
2
3
4
5
6
7
8
9
19
FREQUENCY
20.000GHZ
20.050
20.1
20.15
20.2
POLARIZATION =LHC
18
19
20
21
22
23
24
25
37
FREQUENCY
20.0125GHZ
20.0625
20.1125
20.1625

FIG. 11

【図１２】

FIG. 12


INPUT
FROM ANTENNA FEED ELEMENT
(1200)
(1201)
(1202)
(1203)
(1204)
(1205)
(1206)
FROM FIRST L.O. COMMON TO ALL CHANNELS
1207
1208
COS
SIN
FROM COMMON 2nd L.O.
CROSS-OVER NETWORK
1209
1215
1216
1217
1218
K-BAND I-VCO
1210
L BAND Q-VCO
1211
(1212)
DUAL A/D CONVERTOR
1 1.6 MB/S
2
3
37
FROM OTHER CHANNELS
(1213)
DIGITAL MULTIPLEXER
≈ 60 MB/S
(1214)
HIGH SPEED DIGITAL TRANSMITTER
K-BAND
TO HUB

【図１３】

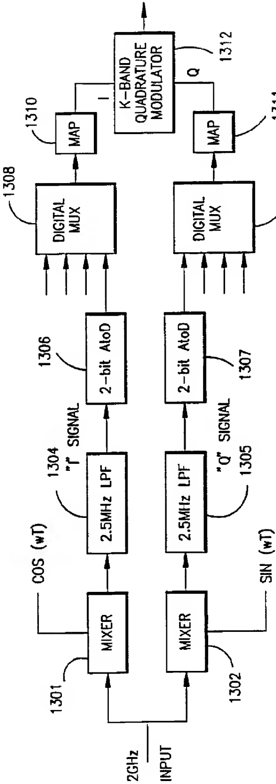

2GHz
INPUT
1301
1302
MIXER
MIXER
COS (wT)
SIN (wT)
2.5MHz LPF
2.5MHz LPF
1304
1305
"I" SIGNAL
"Q" SIGNAL
2-bit AtoD
2-bit AtoD
1306
1307
DIGITAL MUX
DIGITAL MUX
1308
1309
MAP
MAP
1310
1311
I
Q
K-BAND QUADRATURE MODULATOR
1312

FIG. 13

【図１４】

MAPPING OF 2-BIT MULTIPLEXED I AND Q SIGNALS TO K-BAND CARRIER VECTOR

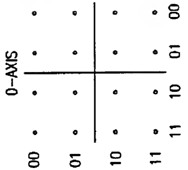


FIG. 14

【図１５】

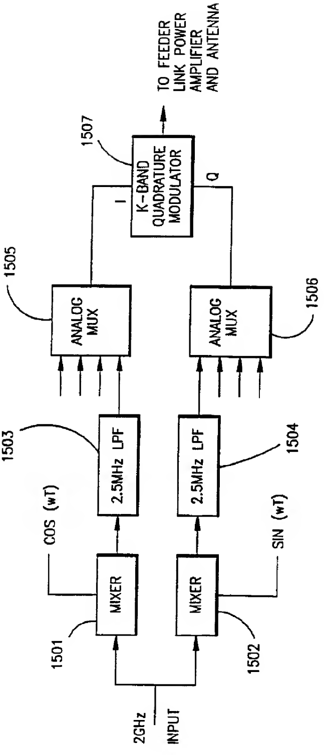

TO FEEDER LINK POWER AMPLIFIER AND ANTENNA
1507
K-BAND QUADRATURE MODULATOR
I
Q
1505
ANALOG MUX
1506
ANALOG MUX
1503
2.5MHz LPF
1504
2.5MHz LPF
COS (wT)
SIN (wT)
MIXER
MIXER
1501
1502
2GHz INPUT

FIG. 15

【図１６】

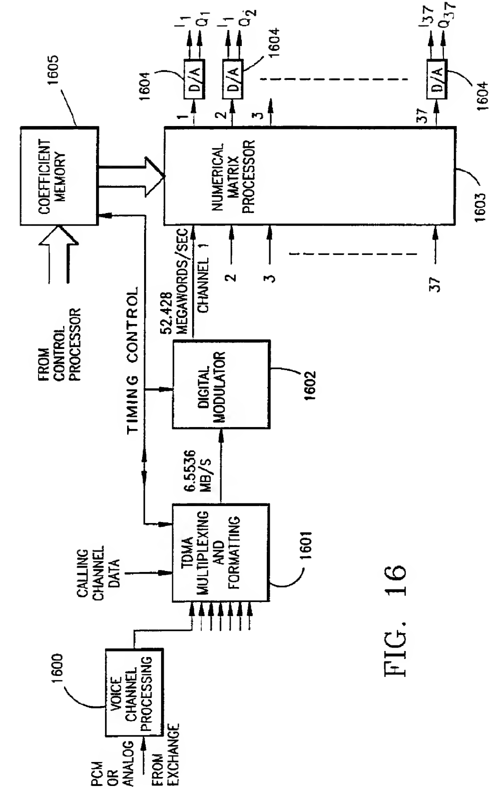

COEFFICIENT MEMORY
1605
FROM CONTROL PROCESSOR
NUMERICAL MATRIX PROCESSOR
1603
D/A
1604
I1
Q1
I1
Q2
I37
Q37
1
2
3
37
TIMING CONTROL
52.428 MEGAWORDS/SEC
CHANNEL 1
DIGITAL MODULATOR
1602
6.5536 MB/S
TDMA MULTIPLEXING AND FORMATTING
1601
CALLING CHANNEL DATA
VOICE CHANNEL PROCESSING
1600
PCM OR ANALOG
FROM EXCHANGE

FIG. 16

【図１７】

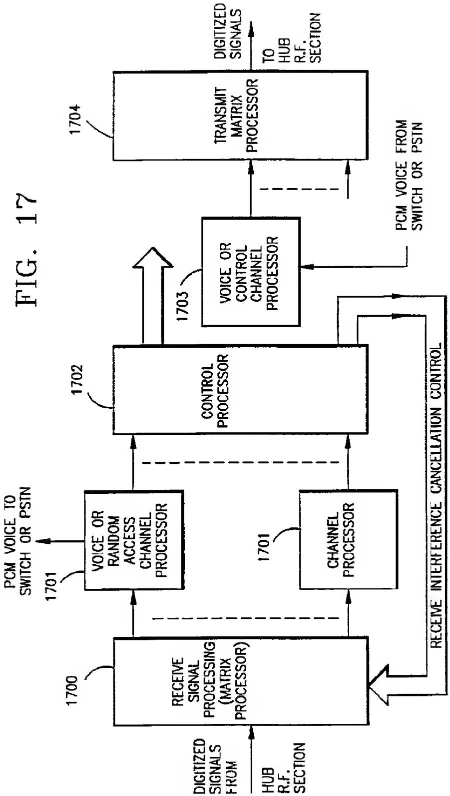

DIGITIZED SIGNALS FROM HUB R.F. SECTION
RECEIVE SIGNAL PROCESSING (MATRIX PROCESSOR)
1700
1701
VOICE OR RANDOM ACCESS CHANNEL PROCESSOR
PCM VOICE TO SWITCH OR PSTN
1701
CHANNEL PROCESSOR
1702
CONTROL PROCESSOR
RECEIVE INTERFERENCE CANCELLATION CONTROL
1703
VOICE OR CONTROL CHANNEL PROCESSOR
PCM VOICE FROM SWITCH OR PSTN
1704
TRANSMIT MATRIX PROCESSOR
DIGITIZED SIGNALS
TO HUB R.F. SECTION

FIG. 17

【図１８】

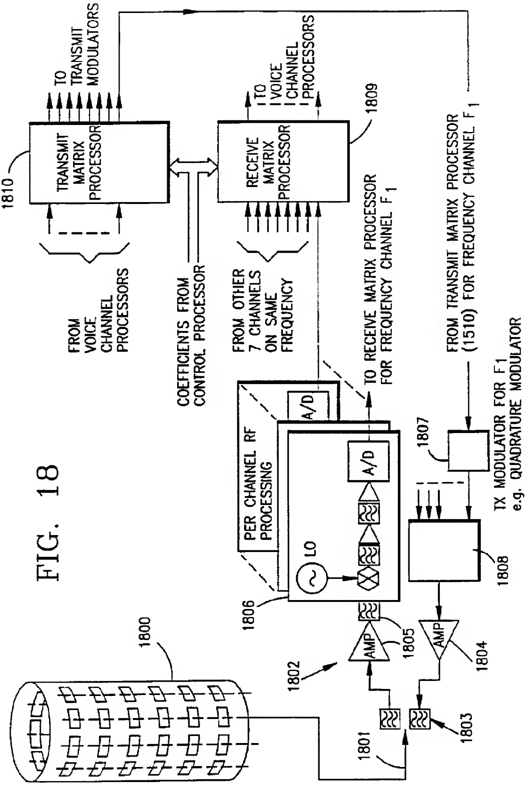

TO TRANSMIT MODULATORS
TRANSMIT MATRIX PROCESSOR
1810
FROM VOICE CHANNEL PROCESSORS
COEFFICIENTS FROM CONTROL PROCESSOR
RECEIVE MATRIX PROCESSOR
1809
TO VOICE CHANNEL PROCESSORS
FROM OTHER 7 CHANNELS ON SAME FREQUENCY
TO RECEIVE MATRIX PROCESSOR FOR FREQUENCY CHANNEL F1
FROM TRANSMIT MATRIX PROCESSOR (1510) FOR FREQUENCY CHANNEL F1
PER CHANNEL RF PROCESSING
A/D
A/D
LO
1806
1807
1808
TX MODULATOR FOR F1 e.g. QUADRATURE MODULATOR
1800
1801
1802
1803
1804
1805
AMP
AMP

FIG. 18

【図19】

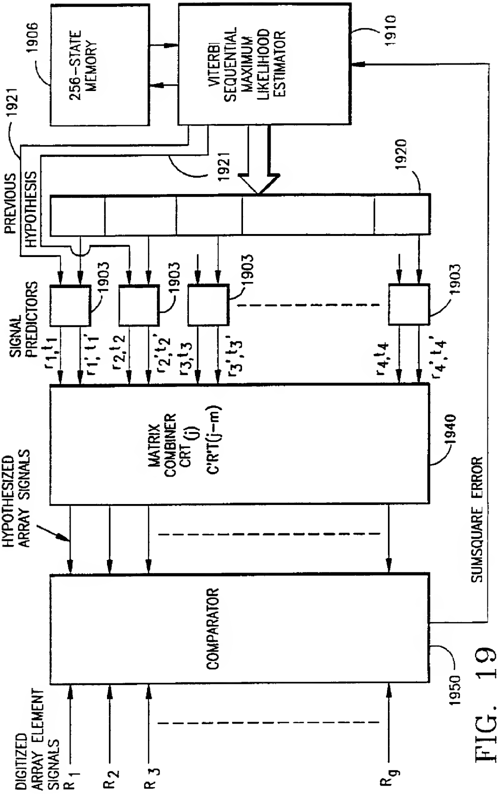

256-STATE MEMORY
1906
VITERBI SEQUENTIAL MAXIMUM LIKELIHOOD ESTIMATOR
1910
1921
PREVIOUS HYPOTHESIS
1920
SIGNAL PREDICTORS
1903
r1,t1
r1',t1'
r2,t2
r2',t2'
r3,t3
r3',t3'
r4,t4
r4',t4'
MATRIX COMBINER CRT(j) C'R'T'(j-m)
1940
HYPOTHESIZED ARRAY SIGNALS
COMPARATOR
1950
SUMSQUARE ERROR
DIGITIZED ARRAY ELEMENT SIGNALS
R1
R2
R3
Rg

FIG. 19

【図２０】

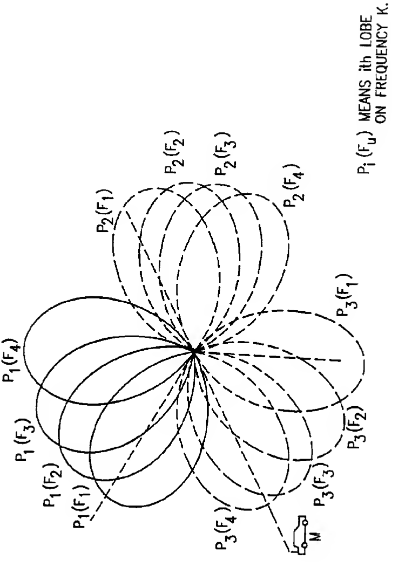

$P_1(F_4)$
$P_1(F_3)$
$P_1(F_2)$
$P_1(F_1)$
$P_2(F_1)$
$P_2(F_2)$
$P_2(F_3)$
$P_2(F_4)$
$P_3(F_1)$
$P_3(F_2)$
$P_3(F_3)$
$P_3(F_4)$
M
$P_i(F_u)$ MEANS ith LOBE ON FREQUENCY K.

FIG. 20

【図２１】

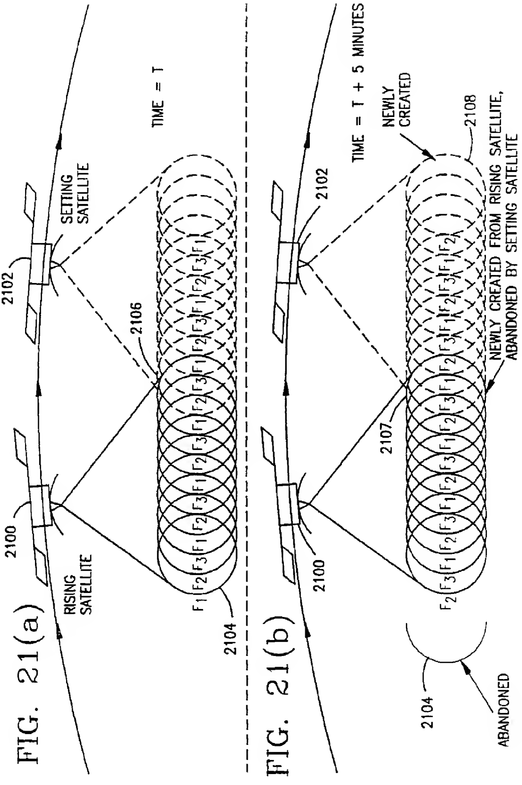


FIG. 21(a)

FIG. 21(b)

【図２２】

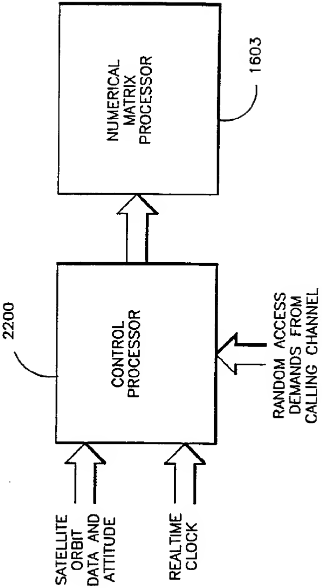

NUMERICAL MATRIX PROCESSOR
1603
CONTROL PROCESSOR
2200
RANDOM ACCESS DEMANDS FROM CALLING CHANNEL
SATELLITE ORBIT DATA AND ATTITUDE
REALTIME CLOCK

FIG. 22

【図２３】

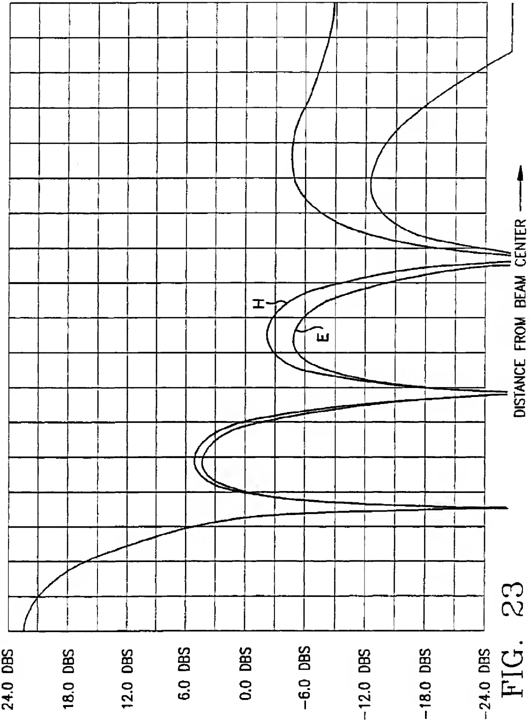

24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
-6.0 DBS
-12.0 DBS
-18.0 DBS
-24.0 DBS
H
E
DISTANCE FROM BEAM CENTER

FIG. 23

【図２４】

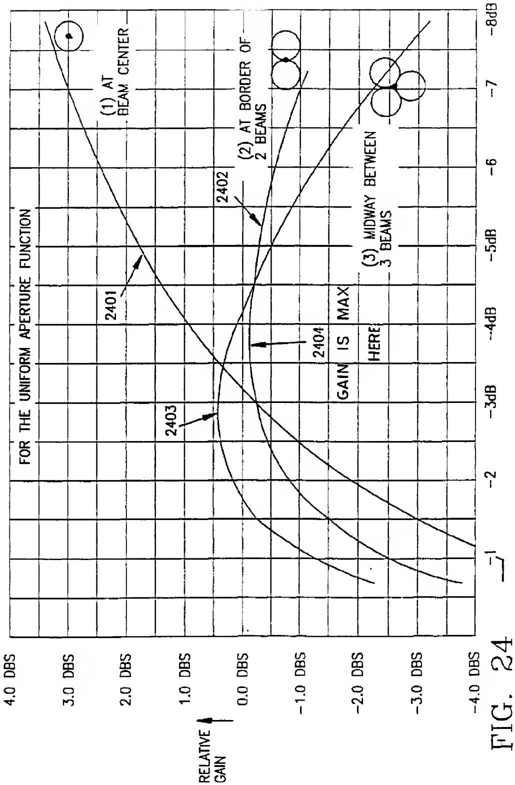

FOR THE UNIFORM APERTURE FUNCTION
(1) AT BEAM CENTER
(2) AT BORDER OF 2 BEAMS
(3) MIDWAY BETWEEN 3 BEAMS
2401
2402
2403
2404
GAIN IS MAX HERE
4.0 DBS
3.0 DBS
2.0 DBS
1.0 DBS
0.0 DBS
−1.0 DBS
−2.0 DBS
−3.0 DBS
−4.0 DBS
RELATIVE GAIN
−1
−2
−3dB
−4dB
−5dB
−6
−7
−8dB

FIG. 24

【図２５】

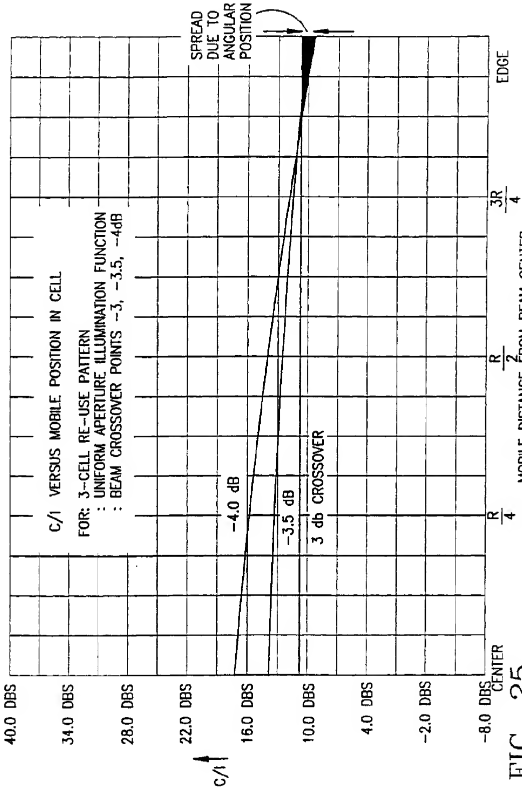

C/I VERSUS MOBILE POSITION IN CELL
FOR: 3-CELL RE-USE PATTERN
: UNIFORM APERTURE ILLUMINATION FUNCTION
: BEAM CROSSOVER POINTS -3, -3.5, -4dB
SPREAD DUE TO ANGULAR POSITION
-4.0 dB
-3.5 dB
3 db CROSSOVER
40.0 DBS
34.0 DBS
28.0 DBS
22.0 DBS
16.0 DBS
10.0 DBS
4.0 DBS
-2.0 DBS
-8.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 25

【図２６】

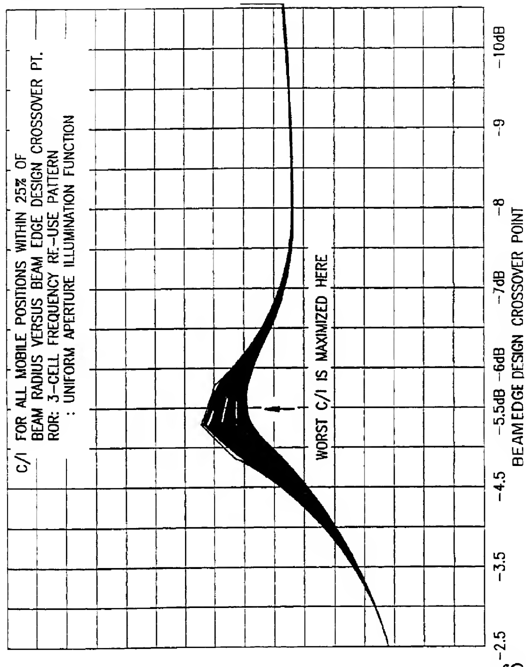

C/I FOR ALL MOBILE POSITIONS WITHIN 25% OF
BEAM RADIUS VERSUS BEAM EDGE DESIGN CROSSOVER PT.
ROR: 3-CELL FREQUENCY RE.-USE PATTERN
: UNIFORM APERTURE ILLUMINATION FUNCTION
WORST C/I IS MAXIMIZED HERE
48.0 DBS
42.0 DBS
36.0 DBS
30.0 DBS
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
C/I
-2.5
-3.5
-4.5
-5.5dB
-6dB
-7dB
-8
-9
-10dB
BEAMEDGE DESIGN CROSSOVER POINT

FIG. 26

【図２７】

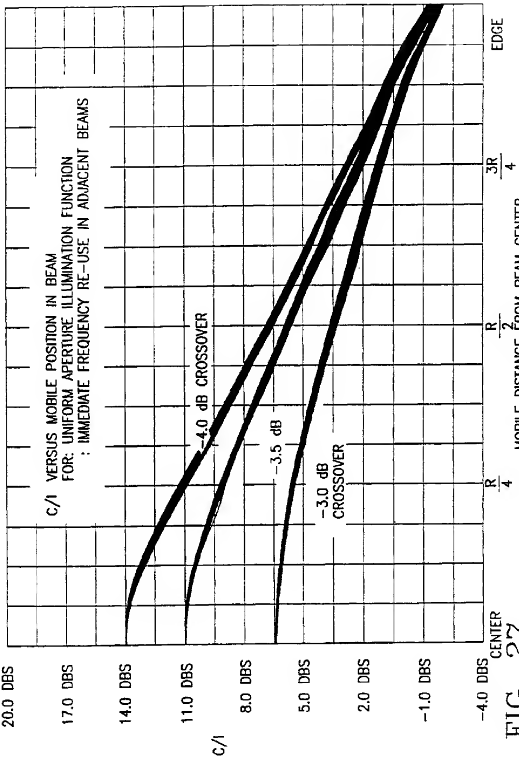

C/I VERSUS MOBILE POSITION IN BEAM
FOR: UNIFORM APERTURE ILLUMINATION FUNCTION
: IMMEDIATE FREQUENCY RE-USE IN ADJACENT BEAMS
-4.0 dB CROSSOVER
-3.5 dB
-3.0 dB
CROSSOVER
20.0 DBS
17.0 DBS
14.0 DBS
11.0 DBS
8.0 DBS
5.0 DBS
2.0 DBS
-1.0 DBS
-4.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 27

【図２８】

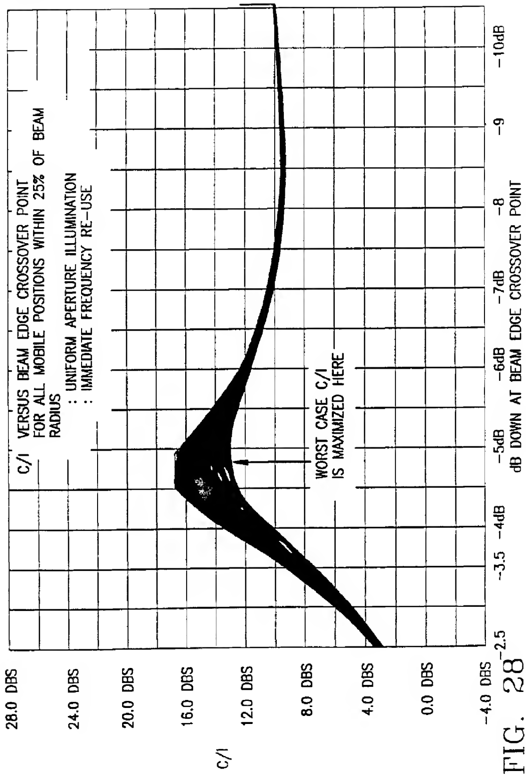

C/I VERSUS BEAM EDGE CROSSOVER POINT
FOR ALL MOBILE POSITIONS WITHIN 25% OF BEAM
RADIUS
: UNIFORM APERTURE ILLUMINATION
: IMMEDIATE FREQUENCY RE-USE
WORST CASE C/I
IS MAXIMIZED HERE
28.0 DBS
24.0 DBS
20.0 DBS
16.0 DBS
12.0 DBS
8.0 DBS
4.0 DBS
0.0 DBS
-4.0 DBS
C/I
-2.5
-3.5
-4dB
-5dB
-6dB
-7dB
-8
-9
-10dB
dB DOWN AT BEAM EDGE CROSSOVER POINT

FIG. 28

【図２９】

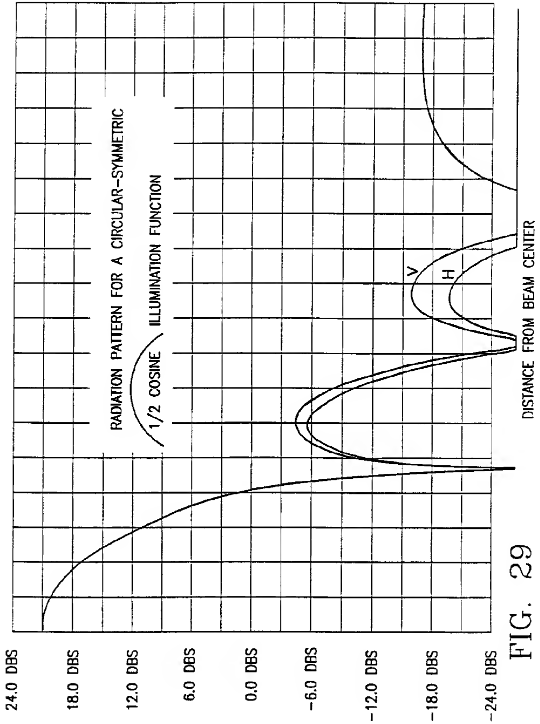

RADIATION PATTERN FOR A CIRCULAR-SYMMETRIC
1/2 COSINE
ILLUMINATION FUNCTION
V
H
DISTANCE FROM BEAM CENTER
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
-6.0 DBS
-12.0 DBS
-18.0 DBS
-24.0 DBS

FIG. 29

【図３０】

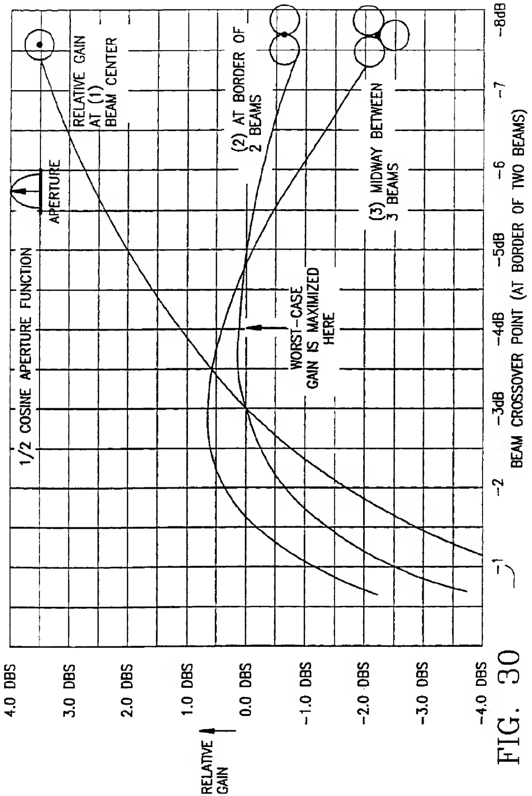

1/2 COSINE APERTURE FUNCTION
APERTURE
RELATIVE GAIN
AT (1)
BEAM CENTER
(2) AT BORDER OF
2 BEAMS
(3) MIDWAY BETWEEN
3 BEAMS
WORST-CASE
GAIN IS MAXIMIZED
HERE
4.0 DBS
3.0 DBS
2.0 DBS
1.0 DBS
0.0 DBS
-1.0 DBS
-2.0 DBS
-3.0 DBS
-4.0 DBS
RELATIVE
GAIN
-1
-2
-3dB
-4dB
-5dB
-6
-7
-8dB
BEAM CROSSOVER POINT (AT BORDER OF TWO BEAMS)

FIG. 30

【図３１】

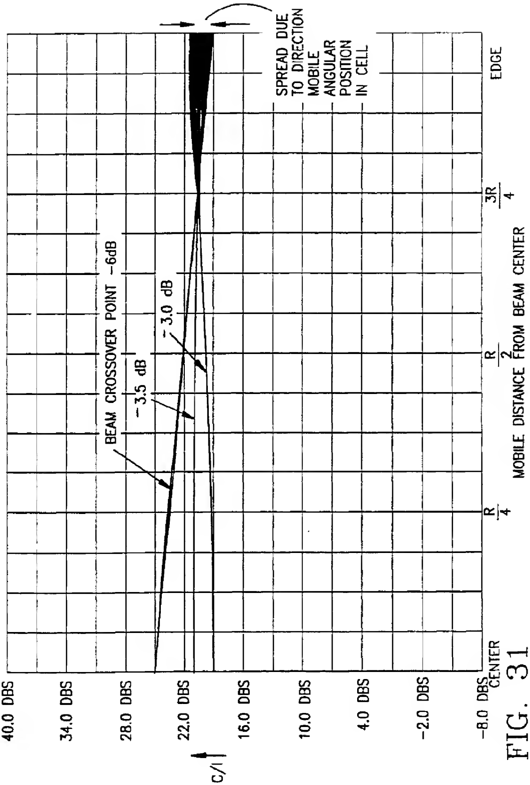

SPREAD DUE TO DIRECTION MOBILE ANGULAR POSITION IN CELL
BEAM CROSSOVER POINT −6dB
−3.5 dB
−3.0 dB
40.0 DBS
34.0 DBS
28.0 DBS
22.0 DBS
16.0 DBS
10.0 DBS
4.0 DBS
−2.0 DBS
−8.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 31

【図３２】

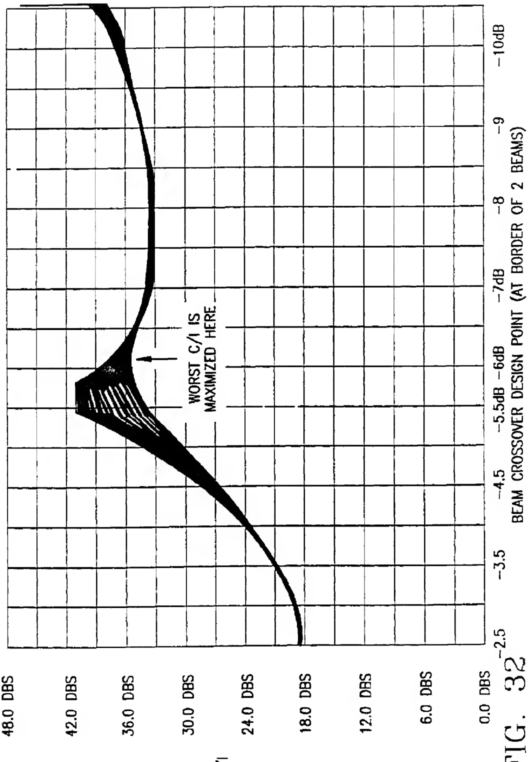

48.0 DBS
42.0 DBS
36.0 DBS
30.0 DBS
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
C/I
-2.5
-3.5
-4.5
-5.5dB
-6dB
-7dB
-8
-9
-10dB
BEAM CROSSOVER DESIGN POINT (AT BORDER OF 2 BEAMS)
WORST C/I IS MAXIMIZED HERE

FIG. 32

【図３３】

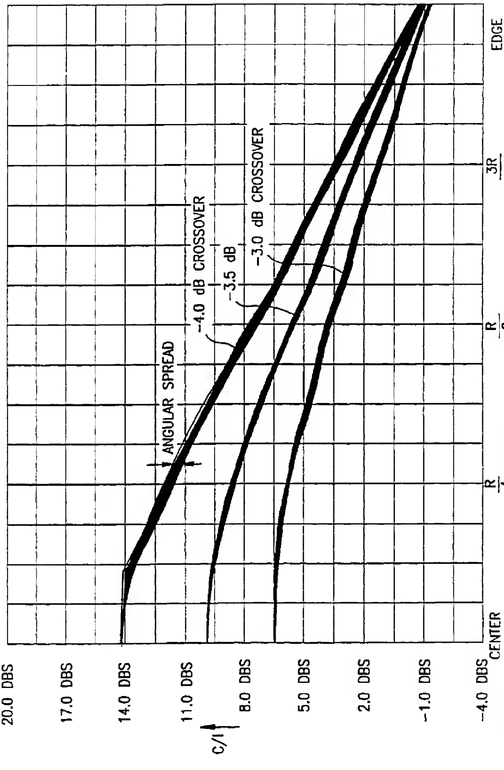

20.0 DBS
17.0 DBS
14.0 DBS
11.0 DBS
8.0 DBS
5.0 DBS
2.0 DBS
-1.0 DBS
-4.0 DBS
C/I
ANGULAR SPREAD
-4.0 dB CROSSOVER
-3.5 dB
-3.0 dB CROSSOVER
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 33

【図３４】

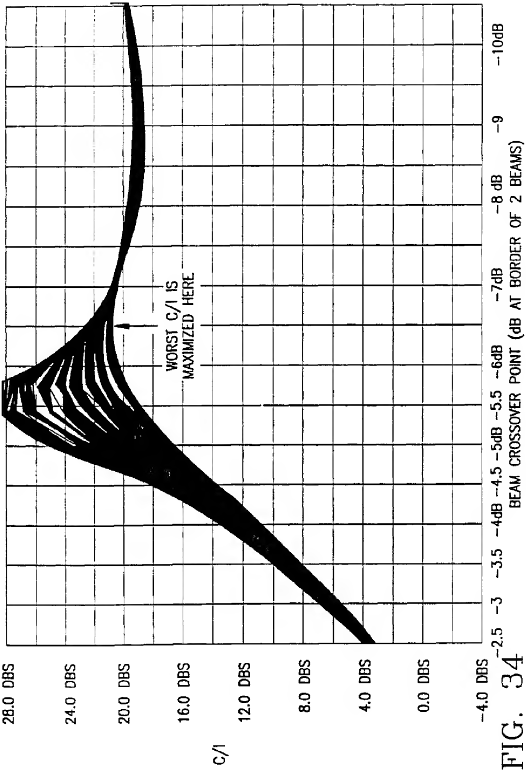
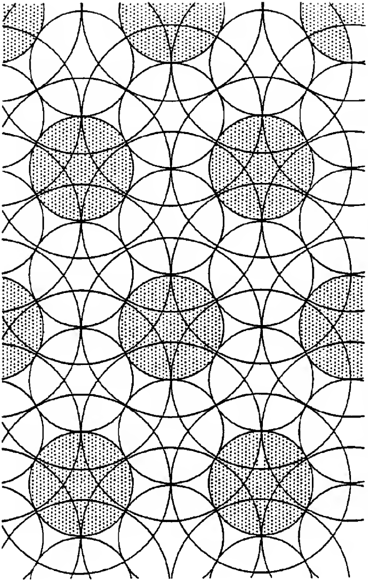

28.0 DBS
24.0 DBS
20.0 DBS
16.0 DBS
12.0 DBS
8.0 DBS
4.0 DBS
0.0 DBS
−4.0 DBS
C/I
−2.5 −3 −3.5 −4dB −4.5 −5dB −5.5 −6dB −7dB −8 dB −9 −10dB
BEAM CROSSOVER POINT (dB AT BORDER OF 2 BEAMS)
WORST C/I IS MAXIMIZED HERE

FIG. 34

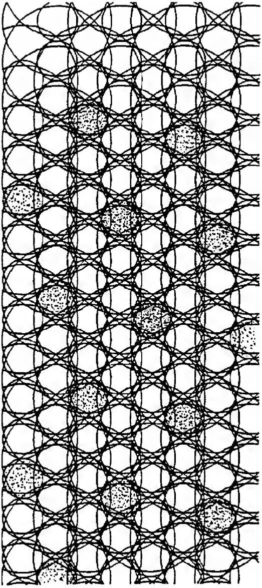
【図３５】



THE CHARACTERS OF EACH GIVEN COLOR ARE TOUCHING (1 CELL PLAN)
BUT THE SMALLER CIRCLES FORM A 3-CELL PATTERN

FIG. 35

【図３６】



7 COLORS USED SYSTEMATICALLY SUCH THAT LARGER CIRCLES OF ANY COLOR TOUCH WHILE THE SMALLER CIRCLES FORM A 7-CELL RE-USE PATTERN WITH THE OTHER COLORS

FIG. 36

【図３７】

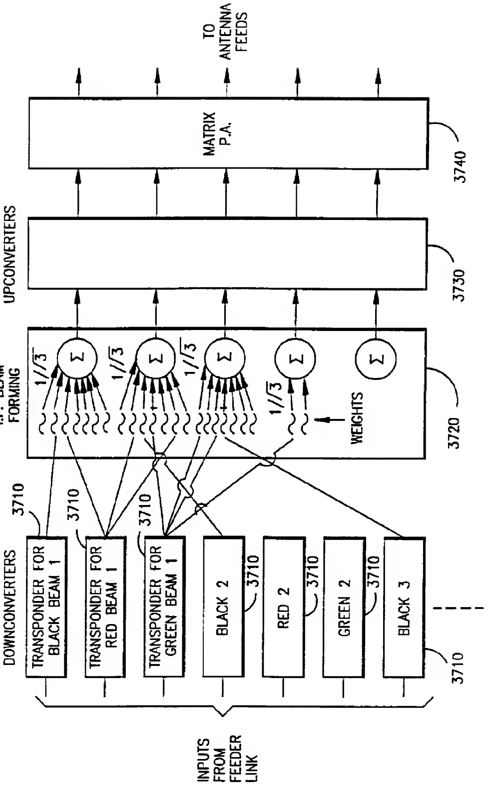
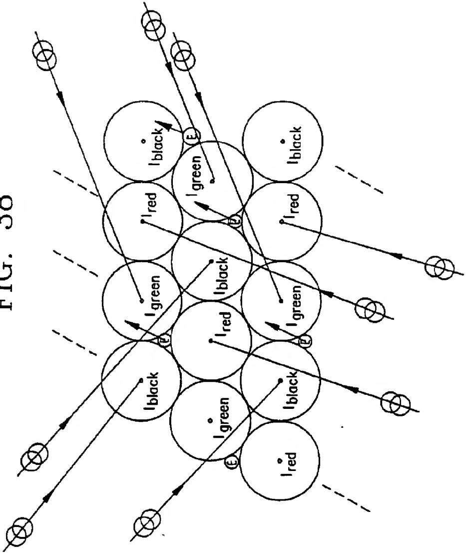

TO ANTENNA FEEDS
MATRIX P.A.
3740
UPCONVERTERS
3730
I.F. BEAM FORMING
1/√3
WEIGHTS
3720
DOWNCONVERTERS
TRANSPONDER FOR BLACK BEAM 1
TRANSPONDER FOR RED BEAM 1
TRANSPONDER FOR GREEN BEAM 1
BLACK 2
RED 2
GREEN 2
BLACK 3
3710
INPUTS FROM FEEDER LINK

FIG. 37

【図３８】

FIG. 38


I_{black}
I_{green}
I_{black}
I_{red}
I_{red}
I_{green}
I_{black}
I_{green}
I_{red}
I_{green}
I_{black}
I_{black}
I_{green}
I_{red}
E
E
E
E
E

【図３９】

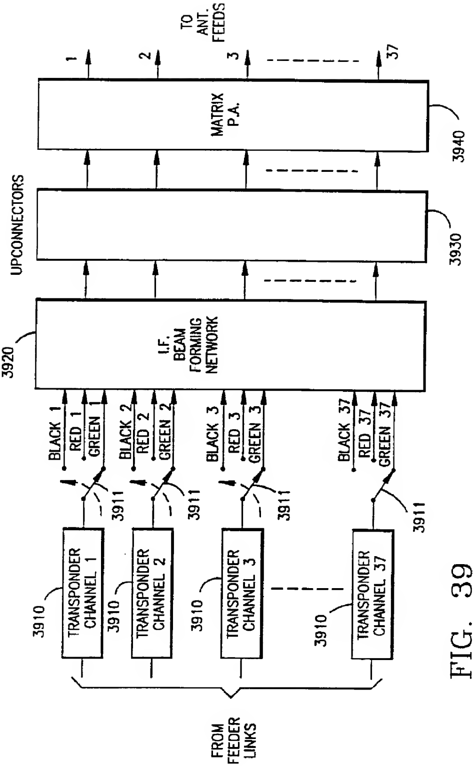

TO ANT. FEEDS
1
2
3
37
MATRIX P.A.
3940
UPCONNECTORS
3930
3920
I.F. BEAM FORMING NETWORK
BLACK 1
RED 1
GREEN 1
BLACK 2
RED 2
GREEN 2
BLACK 3
RED 3
GREEN 3
BLACK 37
RED 37
GREEN 37
3911
3910
TRANSPONDER CHANNEL 1
TRANSPONDER CHANNEL 2
TRANSPONDER CHANNEL 3
TRANSPONDER CHANNEL 37
FROM FEEDER LINKS

FIG. 39

【国際調査報告】

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/00224

A. CLASSIFICATION OF SUBJECT MATTER
IPC(6) :H04J 3/16; H04B 1/04, 1/16, 7/26, 7/185; H04Q 3/04
US CL . 370/95.3, 455/12.1, 13.1, 13.4, 38.1, 54.1
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED
Minimum documentation searched (classification system followed by classification symbols)
U.S. : 370/32, 35, 75, 95.1, 95.3, 104.1, 121, 123; 455/12.1, 13.1, 13.2, 13.4, 17, 22, 38.1, 38.3, 49.1, 53.1, 54.1

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A, E	US, A, 5,398,247 (DELPRAT ET AL) 14 MARCH 1995	1-13, 17-22, 26-30, 44-68, 77-79
A	US, A, 4,907,003 (MARSHALL ET AL) 06 MARCH 1990	14-16, 76
A	US, A, 5,247,702 (SU ET AL) 21 SEPTEMBER 1993	23-25, 31-35, 38
A	US, A, 4,876,737 (WOODWORTH ET AL) 24 OCTOBER 1989	36-37, 39-43
A	US, A, 4,752,967 (BUSTAMANTE ET AL) 21 JUNE 1988	69-75

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

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INTERNATIONAL SEARCH REPORT

International application No.
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Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This international report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. ☐ Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

Please See Extra Sheet.

1. ☒ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest
☐ The additional search fees were accompanied by the applicant's protest.
☒ No protest accompanied the payment of additional search fees.

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/00224

BOX II. OBSERVATIONS WHERE UNITY OF INVENTION WAS LACKING
This ISA found multiple inventions as follows:

This application contains the following inventions or groups of inventions which are not so linked as to form a single inventive concept under PCT Rule 13.1. In order for all inventions to be examined, the appropriate additional examination fees must be paid.

Group I. Claims 1-13, 17-22, 26-30, 44-68 and 77-79, drawn to Time Division Multiple Access, classified in Class 370, subclass 95.3.

Group II. Claims 14-16 and 76, drawn to space satellite with antenna feed network or multiple antenna switching, classified in Class 455, subclass 13.3.

Group III. Claims 23-25, 31-35 and 38, drawn to system with receiver selection with coded sequence, classified in class 455, subclass 38.1.

Group IV. Claims 36-37, 39-43, drawn to satellite transponder transmitting received signals with a new carrier frequency, classified in Class 455, subclass 12.1.

Group V. Claims 69-74, drawn to space satellite with power control, classified in Class 455, subclass 13.4.

Group VI. Claim 75, drawn to wireless communications between a base station and a plurality of mobile stations, classified in Class 455, subclass 54.1.The inventions listed above as Groups do not relate to a single inventive concept under PCTRule13.1because,under PCT Rule 13.2, they lack the same or corresponding special technical features for the following reasons: The inventions listed as Groups I-VI do not relate to a single inventive concept under PCT Rule 13.1 because, under PCT Rule 13.2, they lack the same or corresponding special technical features for the following reasons: Groups I-VI each have different modes of operation, and they have different functions.

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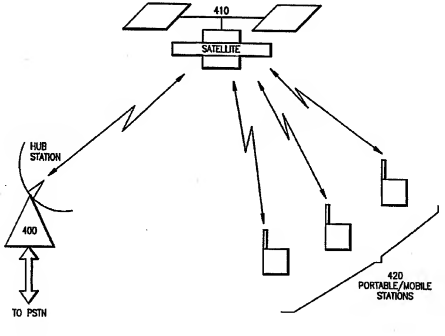

(57) Abstract

Radiocommunication systems and methods for minimizing co-channel interference are disclosed. The systems and methods are applicable to, for example, landmobile communications (420), satellite communication systems (410) and hybrids (400) thereof. Signal processing using matrix models of received and transmitted signals provides for minimized interference.

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A CELLULAR/SATELLITE COMMUNICATIONS SYSTEM WITH IMPROVED FREQUENCY RE-USE

BACKGROUND

The present invention relates to radio communication systems with increased capacity. The system can include a number of roving, automobile-mounted or handheld telephone sets served by either fixed, ground-based stations or by orbiting satellites or by a combination of both. The capacity of such systems to serve a large number of subscribers depends on how much of the radio spectrum is allocated for the service and how efficiently it is used. Efficiency of spectral utilization is measured in units of simultaneous conversations (erlangs) per megahertz per square kilometer. In general, spectral efficiency can be improved more by finding ways to re-use the available bandwidth many times over than by attempting to pack more conversations into the same bandwidth, since narrowing the bandwidth generally results in the need to increase spatial separation between conversations thus negating the gain in capacity. Therefore, it is generally better to increase the bandwidth used for each conversation so that closer frequency re-use is possible.

Spread-spectrum communications systems (e.g., CDMA systems) that increase the signal bandwidths using heavy redundant coding, such that a signal can be read even through interference from other users, offer high spectral efficiency. Using such systems, several users in the same cell can coexist in the same bandwidth, overlapping in both frequency and time. If co-frequency interferers in the same cell can be tolerated, co-frequency interferers one or more cells away can also be tolerated since distance will lessen their interference contribution, so it would be possible to re-use all frequencies in all cells.

Spread-spectrum system capacity is said to be self-interference limited because each unwanted signal that is received simultaneously with the desired signal, and on the same frequency, contributes an interference component. Some systems, however, such as satellite communications systems, are already limited by natural noise, so the wideband spread-spectrum approach is then not necessarily the best technique for maximizing capacity. Consequently it would be desirable to re-use the whole spectrum

in every adjacent cell or region without incurring the self-interference penalty of wideband spread-spectrum.

Figure 1 shows a typical arrangement of a cellular telephone network using land-based stations. This figure is illustrative of such networks only, for example, cells are not always of such regular size and shape and as a general definition a cell may be described as an area illuminated with a distinct signal.

Cells can be illuminated from their geographical centers, but it is more common to illuminate a cluster of three cells from a common site at the junction of the three cells, as site real estate cost is a major economic consideration. The antenna radiation patterns for central illumination of a cell would generally be omnidirectional in azimuth. It is also common to narrow the radiation pattern in the vertical plane so as to concentrate the energy towards land-based telephones and avoid wasting energy skywards. When the transmitters and antennas for three cells are collected onto the same site for economy, the antenna patterns are then only required to illuminate 120 degree sectors, and the resultant azimuthal directive gain largely compensates for the double distance to the far side of the cell. The antenna pattern can be shaped appropriately so as to provide a gain commensurate with the maximum range needed in each direction, which is halved at +/- 60 degrees compared to mid-sector. Thus a sectorized antenna pattern can be narrowed to -12dB at +/- 60 degrees, giving a mid-sector gain of about 8 to 9dB to assist in achieving the maximum range in that direction.

Using central illumination, the U.S AMPS cellular mobile telephone system denies re-use of the same frequency within a 21-cell area around a given cell. This is called a 21-cell frequency re-use pattern and results in co-channel interference being approximately 18dB below a wanted signal when all channels are concurrently in use (commonly called maximum load). Such a 21-cell re-use pattern is illustrated in Figure 2. Certain re-use pattern sizes such as 3, 4, 7 and products thereof (e.g., 9, 12, 21..) result in co-channel interferers being equidistant from the wanted signal and located on the vertices of a hexagon, separated by a number of cells equal to the square root of the pattern size.

In practice, illumination takes place from sites at the junction of three cells. Although the re-use pattern is a 21-cell pattern, it can also be described as 7-sites each having a 3-frequency re-use pattern around the three, 120 degree sectors. The signal to co-channel interference characteristics arising from this form of illumination are not exactly equivalent to those characteristics which result from central illumination (due to the antenna directivity it can be shown that interference with respect to a particular signal arises principally from two other sites whose antennas are firing in the right direction, and not from six equidistant cells which transmit on a common frequency as would be the case in central illumination).

The 3-sector, 7-site method of illumination is sometimes called sectorization, which can give the erroneous impression that an originally larger cell was split into three smaller cells or sectors by use of directional antennas. This impression, however, is inaccurate because the arrangement used for illuminating three cells from the same site is merely an economic arrangement that actually has slight disadvantages over central illumination with respect to technical performance but is otherwise very similar.

Cell-splitting is another concept entirely, being a way of obtaining more capacity per square kilometer by providing base stations more densely on the ground. Introducing cell splitting in an already existing system usually requires complete revamping of the frequency re-use plan, as it is conventionally not possible simply to split a cell, for example, into three cells and to re-use the original frequencies three times over. This would result in the three new cells operating on the same frequency with no spatial separation, which would present a problem for a mobile phone on the boundary between two cells where it receives equal strength (but different content) signals on the same frequency from both. Thus, it would be desirable to allow a cell to be split into sectors with the same frequencies being used in each without the above-described interference problem.

Similar capacity issues arise in designing a satellite communications system to serve mobile or handheld phones. On handheld phones, omnidirectional antennas of indifferent performance are all that in practice the majority of consumers are willing to accept. Directional antennas that have to be oriented toward the satellite or larger,

more cumbersome antennas do not now find favor in the marketplace, so it is necessary for the satellite to provide a high enough signal strength at the ground to communicate with such devices. The signal strength received at the ground from a satellite is usually measured in units of watts per square meter or dBW per square meter on a logarithmic scale. For example, a flux density of the order on -123dBW per square meter is used for voice communication to provide an adequate link margin for multipath fading, shadowing, polarization mismatch etc., using a downlink frequency of 2GHz. The total number of watts radiated by the satellite is then equal to this required flux density times the area of the geographical region it illuminates. For example, to provide such a voice channel anywhere in the entire United States, having an area of 9 million square kilometers requires a total radiated power of:

$$10^{-12.3} \quad x \quad 9x10^{12} \quad = \quad 4.5 \text{ watts from the satellite.}$$

One voice channel would not, of course, provide a useful capacity. Five to ten thousand Erlangs is a more reasonable target for serving the United States. One way of increasing the capacity would be to generate 4.5 watts on other frequencies too, each of which could carry one voice channel; but a 45k watt satellite would be very large and expensive to launch and would not be an economic way to provide 10000 erlangs capacity. It is therefore more efficient, having used 4.5 watts of satellite RF power to create one voice channel's worth of flux density at all places in the United States, to find ways which will allow the voice carried by that flux to be different at different places, thus supporting many different conversations using no more power or bandwidth.

The ability of a satellite to modulate the same radiated flux density differently in different directions depends on the angular discrimination provided by its antenna aperture. The angular discrimination of an antenna (in radians) is on the order of the ratio of the wavelength to the diameter of the antenna. Using an exemplary downlink frequency of 2GHz (15cm wavelength) an antenna of 1.5 meters in diameter theoretically has an angular discrimination on the order of 1/10th of a radian or 5.7 degrees, which, from an orbital height of, for example, 10000 kilometers, allows

discrimination between 37 different directions within the United States coverage area. Thus, the same 4.5 watts of satellite radiated power could then support not just one, but 37 different conversations.

One way of creating 37 different beams is shown in Figure 3. A parabolic reflector focuses the radio energy from a pattern of 37 different feeds down to the earth. An image of the feeds is projected onto the ground forming the desired separately illuminated areas. Unfortunately, using this technique there is spillover from one area to another, and in any case a mobile phone on the boundary between two or three cells receives equal signals from two or three feeds. If these signals are independently modulated, the phone receives a jumble of three conversations which it cannot decipher. Accordingly, conventional systems have been unable to exploit the potential capacity increases which would be realized using discrimination.

SUMMARY

These and other drawbacks and difficulties found in conventional radio communication systems, satellite communication systems and hybrids thereof are overcome according to the present invention.

According to exemplary embodiments of the present invention, matrix processing can be used to form numerical combinations of data sample streams. The matrix coefficients are selected, and can be periodically adjusted, so that each of a plurality of receivers receives its intended signal with substantially zero interference.

According to another exemplary embodiment of the present invention, signal processing does not adapt to the movement of mobile phones or to new call set-up and termination, but operates in a deterministic way and instead the traffic is adapted to the deterministic characteristics of the signal processing using a dynamic traffic channel assignment algorithm.

BRIEF DESCRIPTION OF THE DRAWINGS

The foregoing, and other, objects, features and advantages of the present invention will be more readily understood upon reading the following detailed description in conjunction with the drawings in which:

Figure 1 illustrates a conventional land based cellular network;

Figure 2 illustrates a conventional 21-cell frequency re-use plan;

Figure 3 shows a conventional satellite implementation of 37 beams illuminating a region of the Earth;

Figure 4 illustrates an illumination pattern used to describe a feature of the present invention;

Figure 5 shows a 3-cell frequency reuse plan;

Figure 6 shows a satellite-mobile communication system according to an exemplary embodiment of the present invention;

Figure 7 illustrates a mobile-to-hub transponder according to an exemplary embodiment of the present invention;

Figure 8(a) illustrates a hub-to-mobile satellite transponder according to an exemplary embodiment of the present invention;

Figure 8(b) illustrates a combining network for a power amplifier matrix according to another exemplary embodiment of the present invention;

Figure 9 shows a hubstation according to an exemplary FDMA embodiment of the present invention;

Figure 10 illustrates coherent beam signal transmission according to an exemplary embodiment of the present invention;

Figure 11 shows spectral characteristics using dual polarizations on k-bank hublinks according to an exemplary embodiment;

Figure 12 is a block diagram illustrating phase coherent transportation of the beam signals according to an exemplary embodiment;

Figure 13 is a block diagram illustrating phase coherent transportation of beam signals according to another exemplary embodiment of the present invention;

Figure 14 illustrates mapping of 2-bit multiplexed I and Q signals to a K-band carrier vector;

Figure 15 is a block diagram illustrating yet another exemplary embodiment of phase coherent beam signal transportation;

Figure 16 is a block diagram illustrating hubstation transmit signal processing according to an exemplary TDMA embodiment of the present invention;

Figure 17 illustrates connections between a receive control processor and a transmit control processor according to an exemplary embodiment of the present invention;

Figure 18 shows a land-cellular exemplary embodiment of the present invention;

Figure 19 is a block diagram illustrating the maximum likelihood demodulation of signals from an antenna array according to an exemplary embodiment of the present invention;

Figure 20 shows an exemplary arrangement of staggered sector patterns;

Figures 21(a) and 21(b) illustrate progressive illumination patterns according to an exemplary embodiment of the present invention;

Figure 22 is a block diagram illustrating part of an exemplary implementation of a dynamic channel assignment embodiment of the present invention;

Figure 23 is a graphical representation of an exemplary radiation pattern for a circular-symmetric, uniform aperture illumination function;

Figure 24 is an exemplary graph of relative signal gain versus beam crossover points;

Figure 25 is an exemplary graph illustrating C/I versus mobile position in cell for a 3-cell frequency re-use pattern;

Figure 26 is an exemplary graph illustrating C/I versus beam edge crossover point;

Figure 27 is an exemplary graph illustrating C/I versus mobile position in cell for an immediate frequency re-use system;

Figure 28 is an exemplary graph illustrating C/I versus beam edge crossover point for an immediate frequency re-use system;

Figure 29 illustrates an exemplary radiation pattern for a circular-symmetric, ½ cosine aperture illumination function;

Figure 30 is an exemplary graph of relative signal gain versus beam crossover points for the illumination function pattern of Figure 29;

Figure 31 is an exemplary graph illustrating C/I versus mobile position in cell for a 3-cell re-use pattern for the illumination function of Figure 29;

Figure 32 is an exemplary graph illustrating C/I versus beam crossover point at all mobile positions within 25% of cell radius for a 3-cell re-use pattern for the illumination function of Figure 29;

Figure 33 is an exemplary graph illustrating C/I versus mobile position in cell for immediate frequency re-use for the illumination function of Figure 29;

Figure 34 is an exemplary graph illustrating C/I at all points within 25% of beam radius as a function of dB down at beam edge crossover for an immediate frequency re-use system using the aperture illumination function of Figure 29;

Figure 35 illustrates beam and cell patterns according to an exemplary embodiment of the present invention;

Figure 36 illustrates another exemplary beam and cell pattern using seven communication channels;

Figure 37 is a block diagram of a fixed beam forming apparatus according to yet another exemplary embodiment of the present invention;

Figure 38 is a diagram of current injection and extraction points illustrating a beam forming apparatus according to an exemplary embodiment of the present invention; and

Figure 39 illustrates an exemplary TDMA embodiment of the beam forming apparatus of Figure 37.

DETAILED DESCRIPTION

Initially, it is helpful to understand the interference problems associated with the transmission of signals from conventional antenna arrays, such as the one illustrated in Figure 3. Figure 4 illustrates a cross section of the illumination intensity produced on the ground from an antenna, such as the antenna shown in Figure 3. Even for a best interference case where a mobile unit is located at the center of beam 2 (point A) the illumination from beams 1 and 3 is not zero, but only somewhat reduced. The total signal received by mobile 2 can be described as the sum of three components, such as:

an amount C21 times the beam 1 signal S1 (small)
an amount C22 times the beam 2 signal S2 (large)
an amount C23 times the beam 3 signal S3 (small)

Considering now the reverse (uplink) direction, and assuming reciprocal propagation, the satellite receives in beam 2 a contribution from three mobiles, namely C21.M1 + C22.M2 + C23.M3 where M1, M2, M3 are respectively the signals radiated from mobiles in cells 1, 2 and 3. If mobile 1 is not close to the edge of beam 2, C21 will be small; since mobile 2 is within beam 2, C22 will be large; and if mobile 3 is not close to the edge of beam 2, C23 will be small. Thus as long as mobiles are ideally placed and not on the edges of cells, it may be that the level of inter-cell interference can be tolerated.

On the other hand, if a mobile unit is, for example, close to the boundary between cell 1 and cell 2, the coefficient C21 will be large and Ml will interfere with the decoding of signal M2. The conventional technique for avoiding this problem is to deny use of the same frequency in immediately adjacent cells. For example, the 3-cell frequency re-use pattern shown in Figure 5 might be used. The shaded cells in Figure 5 are those using a first frequency f1, while the other cells use f2 and f3 in the indicated pattern. It can be seen that cells using the same frequency f1 do not abut, and have edge-to-edge separations of just less than one cell diameter. A mobile on the edge of one beam is quite far down the illumination intensity curve of other cells using the same frequency, and thus avoids interference. However, the drawback is that only one-third of the available frequencies may be used in each cell, reducing the spectral utilization efficiency by a factor of three. Accordingly, the present invention provides, among other features, a means of cancelling the co-channel interference without the loss of spectral efficiency entailed in denying spectral re-use in adjacent cells.

If the expressions for the signals received in all beams B1, B2, B3 ... etc., are collected together, and we assume for the moment the same number of mobile signals as beams, then the following set of equations results:

B1 = C11.M1+C12.M2+C13.M3+C14.M4.....C1n.Mn
B2 = C21.M1+C22.M2+C23.M3+C24.M4.....C2n.Mn
B3 = C31.M1+C32.M2+C33.M3+C34.M4.....C3n.Mn
. . . . . .
Bn = Cn1.M1+Cn2.M2+Cn3.M3+Cn4.M4.....Cnn.Mn

which can be abbreviated to $\underline{B} = C.\underline{M}$ where $\underline{B}$ and $\underline{M}$ are column vectors and C is a square, n x n matrix of coefficients.

From the signals received by the satellite in each of its beams, it is desirable to determine the signals transmitted by the mobiles; according to the present invention this can be done by solving the above set of equations to obtain:

$$\underline{M} = C^{-1}.\underline{B}$$

This solution can be obtained as long as the matrix C is invertible (i.e., has a non-zero determinant) and results in cancellation of substantially all interference between mobile signals and complete separation therebetween. All the elements of the above equations, that is mobile signals Mi, beam signals Bk and matrix elements Cki, are complex numbers having both a real and an imaginary component so as to be able to represent not only signal amplitude differences but also signal phase relationships. According to the present invention the signals received in the different antenna beams are sampled at the same time at a rate sufficient to capture all signal components of interest according to Nyquist's criteria. One set of such samples forms the column vector $\underline{B}$ at any instant, and each such vector is multiplied by the inverse of C, for example, once per sample period to obtain a set of samples M representing interference free mobile signals. Successive values of the same element of M form the sample stream corresponding to one mobile signal. This stream is fed to a digital signal processor for each mobile signal that turns the sample stream into, for example, an analog voice waveform or 64KB PCM digital voice stream as required by the telephone switching system to which the system is connected.

According to another aspect of the present invention, the matrix C does not have to be inverted every sample period, but can be inverted less frequently or only once at the beginning of a call. The matrix C and its inverse vary relatively slowly because the rate at which the C coefficients change due to the mobile unit shifting position within the beams, or due to the beam illumination intensity distributions changing due to satellite movement in the non-geostationary case, is relatively low. In an exemplary satellite embodiment of the present invention, typical cell sizes are in the hundreds of

kilometers diameter range and satellites orbiting at medium altitudes take an hour or two to pass by a typical cell. Thus the need to compute a new matrix inverse due to movements may not arise for the duration of, for example, a typical 3-minute telephone call. The principal reason that changes in the inverse C-matrix would be beneficial, however, is that conversations are continually being connected and disconnected. If n=37, for example, and the average call duration is 3 minutes, then on the average one mobile and its corresponding column of matrix C drops out and is replaced with another column of coefficients every 5 seconds. The process whereby new inverse C matrices are introduced when this occurs will be explained later, suffice it to say that this represents a relatively negligible computational effort compared to the total digital signal processing involved in demodulating and decoding 37 mobile signals.

An exemplary embodiment which applies these principles will now be described with reference to Figures 6-11.

Figure 6 illustrates a plurality of portable stations 420 in communication via satellite 410 with a hubstation 400. The hubstation is connected, for example via a local exchange, to a public switched telephone network (PSTN) to allow calls to be placed between the portable phones and any telephone subscriber worldwide, as well as between the satellite phones. The satellite receives signals from the portable phones at a relatively low microwave frequency, such as 1600 MHz. At such frequencies the transmitters in battery operated phones can be efficient and their antennas can be small and omnidirectional. The satellite translates the received signals from 1600 MHz to a higher frequency for relaying to the hubstation.

A higher frequency can be used because the bandwidth needed on the satellite-to-hub link is at least n times the bandwidth allocated at 1600 MHz for each beam, where n is the number of beams. For example, if 6 MHz of bandwidth is re-used in each of 37 beams at 1600 MHz, then at least 37x6 or 222 MHz of bandwidth will be needed on the satellite-hub link. Since one method of maintaining coherent beam signal transport uses at least twice this bare minimum bandwidth, and the reverse direction requires the same amount, 1 GHz of bandwidth is needed. This suggests that a carrier frequency around, for example, 20 GHz is appropriate for the satellite-hub forward and return links.

At such a frequency, even relatively small hubstation dishes will have very narrow beamwidths, so that exclusive use of this bandwidth by any one system is not necessary, and the entire bandwidth can be re-allocated to other satellites and ground stations without interference as long as the sightline from a first ground station to a first satellite does not intersect with a second satellite. This can be avoided by allocating unique "stations" to satellites in geostationary orbit, or, in the case of lower orbiting satellites that move, the probability of intersection is low and can be handled by having an alternative hub location which is activated when such an event threatens.

Figure 7 shows a block diagram of an exemplary satellite transponder for relaying mobile-originated signals to the hubstation. The L-band (e.g., 1600 MHz) multi-beam satellite antenna 470 receives signals from a plurality of mobile phones distributed between the various beams and amplifies them in respective low-noise amplifiers 480. The composite signal from each beam contains, for example, signals from 400-500 mobile phones using different frequencies spaced at 12.5 KHz intervals over a total bandwidth of 6 MHz. The composite signals of each beam are downconverted in respective mixers 440 to obtain baseband signals, for example, spanning the range of 1-7 MHz. This type of signal will be referred to hereafter as a "video" signal as it is typical of the frequency range spanned by signals from a TV camera. To downconvert the composite received signal to the video signal, the downconverters can, for example, be image-rejection type downconverters. The downconversion process can optionally take place in one or more steps using appropriate intermediate frequencies.

The downconvertors in the system can use the same local oscillator signal so as to preserve the phase relationships at the downconverted frequencies that were received at the antennas. The inadvertent introduction of fixed phase mismatches and small amplitude differences between channels is not a problem as this can be calibrated out by analog or digital processing at the hubstation.

The baseband signals are used to modulate respective carriers at the satellite-hub frequency band, e.g., 20 GHz. If single-sideband modulation of a 1-7 MHz "video" signal were applied to a 20 GHz carrier frequency, the resulting signal would occupy the frequency range 20.001 to 20.007 GHz.

However, using single-sideband modulation can make it difficult to preserve the phase coherency between the beam signals. Accordingly, double-sideband modulation techniques can be used instead. For example, the 1-7 MHz video signal can be used to frequency or phase modulate a 20 Ghz carrier frequency. The frequency range occupied by the modulated carrier would then be approximately 19.993 - 20.007 MHz, or more, depending on the frequency or phase deviation employed. To allow some margin over the bare 14 Mhz bandwidth, a 25 Mhz carrier spacing might be appropriate in the 20 Ghz band. Thus, 37x25 or 925 MHz can be used for the one-way satellite-hub link. This bandwidth can be halved by intelligent use of orthogonal polarizations as described later.

Figure 8(a) shows an exemplary satellite transponder for the hub-mobile relay direction. The same method described above for the mobile-to-hub transmissions can be used in reverse for the coherent transport of multiple beam signals to the satellite. The hub station (not shown) transmits a number of Ka band frequency or phase modulated carriers to the satellite. These are received using a suitable Ka band antenna 360, amplified in a common low-noise amplifier 350, and then fed to FM receiver bank 340 where each carrier is demodulated by a respective receiver to produce a video frequency version of the signals for transmission in respective beams. These video signals, for example occupying the band 1-7 MHz, are then upconverted in respective upconverters 320, using a common local oscillator 330 to preserve relative phase relationships, and then amplified using power amplifier matrix 310 for transmission via multi-beam antenna 300 to the mobile phones. A suitable frequency for the satellite-to-mobile link is, for example, 2.5 GHz (S-band). The amplifiers in the power amplifier matrix can be linear amplifiers to reduce intermodulation between signals destined for different phones. The power amplifier matrix can for example, either be a bank of n separate amplifiers each associated with respective beams, or a bank of N (greater or equal to n) amplifiers coupled by nxN Butler matrices at their inputs and Nxn Butler matrices at their outputs. The effect of the Butler matrices is to use each amplifier to amplify part of every beam signal, thus evening the load, providing graceful degradation in the event of failure, and reducing intermodulation by absorbing a proportion of the intermodulation energy in N-n dummy loads. Examples of such power amplifier

matrixes can be found in U.S. Patent Application Serial No. ________________ entitled "Waste Energy Control and Management in Power Amplifiers" and filed on January 11, 1994 which is incorporated here by reference.

According to another exemplary embodiment of the present invention, in communication systems using TDMA signals relayed through an earth-orbiting satellite having a communications transponder using such a matrix power amplifier, the power amplifier can have its input Butler combining network located at the ground station instead of the satellite. A Butler combining operation may be performed by digital signal processing at the ground station to form weighted sums of the desired beam signals to generate drive signals corresponding to each amplifier of the matrix power amplifier. These weighted sums are transmitted using coherent feeder links to the satellite's communications transponder which receives them and translates them to a second frequency band for driving the power amplifier in such a way that, after Butler combining the power amplifier outputs, the output signals correspond to signals desired to be transmitted in different antenna beam directions to respective ground-terminals, which may be, for example, a small handportable station.

The resulting satellite circuitry is shown in Figure 8(a). Note that the input combiner which is normally present has been omitted since this function is now performed at the ground station, as illustrated by the dashed rectangular outline 800. The antenna 810, signal processing including linear amplifier 820, feeder link receivers and downconverters 830 and output combiner 840 can be implemented in the conventional manner and thus are not further described herein.

This embodiment may be advantageous for certain situations, for example, dynamic reallocation of power between antenna beams and timeslots may be accomplished without large variations in the corresponding forward feeder link signals, because each feeder link carries part of every beamsignal instead of all of one beam signal. Additionally, pre-distortion of signals sent on the forward feeder links may be applied to further compensate for distortion in the associated transponder channel power amplifiers. Moreover, in the case of the over-dimensioned matrix power amplifier described in the above-incorporated "Waste Energy Control and Management in Power Amplifiers" application, the number of feeder links is greater than the number of

independent beam signals to be created, thus affording a measure of redundancy against failure.

Figure 9 shows a block diagram of a hubstation according to an exemplary embodiment of the invention. The hub antenna 600 receives Ka band carriers from the satellite and, after common low-noise amplification and optional downconversion in block 610, the signal is divided between a number of receivers for respective Ka-band carriers to obtain the beam signals $B_1 \ldots B_n$. Each beam signal is composed of a multiplicity of voice-modulated channel frequencies which are separated in channel separation filters 630.

The channel separation filters 630 can be analog components such as crystal filters, and may involve frequency conversion of a selected channel frequency to a common lower frequency (e.g., 12.5-25 KHz, or 455 KHz) for A/D conversion. The selected channel signal having been converted to a suitable frequency is A/D converted in A/D convertors 640. An exemplary A/D convertor technique suitable for use at low intermediate frequencies such as 455 KHz is the technique described in U.S. Patent 5,048,059 to Paul W. Dent entitled "Log-Polar Signal Processing", which is incorporated here by reference, which preserves the full complex nature of the signal by simultaneously digitizing its phase and its amplitude. Instantaneous phase can be digitized for example, using the technique described in U.S. Patent 5,084,669 to Paul W. Dent entitled "Direct Phase Frequency Digitization", which is also incorporated here by reference. Phase digitization of all n beam signals corresponding to one channel frequency can be carried out using the technique described therein by repeating certain elements (i.e., the trigger circuits and holding registers) n times and sharing others (i.e., the reference frequency counter) as necessary to preserve relative phase coherency. Alternately, digital filters can be used instead of analog filters if the composite beam signals are first digitized, in which case the A/D converters 640 in Figure 9 would not be needed.

The numerical results of A/D conversion are fed sample by sample to numerical matrix processor 650. There is one such processor per channel frequency, but only the processor for channel frequency (m) is shown for clarity. The matrix processor processes the digitized beam signals to separate out up to n separate mobile phone

transmissions $M_1 \ldots M_n$ and transfer a sample stream corresponding to each mobile phone transmission to voice channel processor 660. The voice channel processor numerically performs demodulation of the signal and error correction decoding and transcoding of digitized voice from the bit rate and format used over the satellite to standard PCM telephone system format for connection via a digital exchange (not shown) to the PSTN. Thus the exemplary structure shown in Figure 9 accomplishes decoding of nxm voice channels, where n is the number of beams and m is the number of frequencies per beam. For example, with n=37 and m=400, the system has a 14800 voice channel capacity potential:

The explanation of Figure 9 relates to a system wherein one voice channel is carried per frequency (i.e., a Frequency Division Multiple Access (FDMA) system). However, the present invention can also be applied to Time Division Multiple Access (TDMA) systems. In TDMA systems, several mobile phone signals are carried on the same channel frequency by dividing a repetitive frame period into time slots, and allocating one time slot in each frame to one mobile phone signal. The exemplary block diagram of Figure 9 can be applied even when the sample streams from A/D convertors 640 represent TDMA signals. However, the matrix processor 650 will now separate a different set of mobile signals in each timeslot, so that the matrix coefficients are now multiplexed between several sets, each of which correspond to a time slot. This can be an economic arrangement because, for a given number of voice channels complexity, the channel filters 630 will be fewer in number by a factor equal to the number of timeslots per carrier, the A/D convertors are correspondingly fewer, the number of matrix processors is reduced correspondingly although each has to operate at a higher input sample rate, and each voice channel processor can sequentially process the signals in consecutive timeslots and thus achieve the same total number of voice channels capacity while economically time-sharing components.

Each numerical matrix processor 650 is shown receiving a control signal. This control signal can be generated by a separate computer (not shown) which controls the connecting and disconnecting of calls to mobile phones, requiring changes to the matrix of coefficients used by the processor for separating out mobile signals from the beams. It was mentioned earlier that this separation can be achieved if the inverse of the C

matrix was not numerically ill-conditioned. If two mobiles are located exactly at the same point on the earth, their two corresponding columns of the C matrix will be identical, which causes the determinant to be zero and the inverse not to exist. Thus, for the C-matrix to be invertible the mobiles shall be spaced far enough apart on the ground. If they approach each other too closely, the C-matrix becomes ill-conditioned.

According to an aspect of the present invention, however, when this situation threatens, one of the two (or more) approaching mobiles changes frequency to a channel where the other mobiles using the same frequency are adequately separated. It is the function of the control computer to determine, at least at call set-up and optionally at intervals thereafter, which of the available channel frequencies is most suitable for allocating to a new mobile, or for handing over an ongoing conversation. If there is no free capacity in a system the system is said to be blocked and subscribers cannot place calls, much to their annoyance. When the system is underloaded there are, at least on some frequencies, fewer mobile signals than beams, thus the matrix C is not square. It will be shown later how the excess degrees of freedom provided in underloaded systems can then be used, not only to separate mobile signals from each other thus avoiding mutual interference, but also to maximize the signal quality received from the worst-case mobile. This solution changes when an extra mobile signal has to be accommodated and the control computer can evaluate in advance the potential impact on signal qualities. Thus a strategy for allocating a channel according to an exemplary embodiment of the present invention is to evaluate the impact on the signal quality corresponding to the worst case mobile on each channel through the inclusion of the new signal in the computations. The channel which suffers the least degradation, or meets the highest quality for the worst-case mobile, is then logically selected as the one to use for the new signal. This results in the group of mobiles assigned to the same frequency being as widely spatially separated as possible.

Figure 10 shows an exemplary arrangement for coherent transmission of "video" signals from each beam. The video signal from the first antenna feed element's (beam) downconvertor (not shown) is fed to the voltage control input of a 20 GHz voltage controlled oscillator (VCO) 1000. The video signal frequency modulates the VCO.

Successive VCO's with their center frequencies offset by the desired channel spacing (e.g., 25 MHz) are used for the signals from antenna feed elements 2,3.... to n/2. The VCO center frequency for signal 2 according to this exemplary embodiment is 25 MHz (i.e., 1 GHz/40) higher than that for signal 1, and the VCO 1001 frequency for signal n/2 is thus (n/2-1) x 1 GHz/40 = (n-2)/80 GHz higher than that for signal 1. The signals from the VCOs are summed in summer 1002 which can be, for example, a waveguide or stripline directional coupler network, and the sum is amplified in a common amplifier 1003 which can, for example, be a travelling wave tube amplifier (TWTA).

A parallel arrangement is used to deal with the other half of the video signals numbered n/2+1 to n. The VCO 1004 for signal n/2+1 is offset by half a channel spacing (i.e., by 12.5 MHz in the above example where channel spacing is 25 MHz) from that of VCO 1000 and this offset is maintained up to VCO 1005 such that the set of frequencies used in the parallel arrangement are all offset half of a channel from those of the first arrangement. This minimizes any interference which may be caused by imperfect polarization isolation in the dual polarization transmission system. The outputs of the two TWTAs are connected to, for example, dual-circular-polarized horn antenna 1009 via a polarizer 1008. The function of the polarizer 1008 is to launch a Right Hand circularly polarized signal into horn antenna 1009 corresponding to the signal from TWTA 1003 and simultaneously a Left Hand Circularly polarized signal corresponding to the signal from TWTA 1007.

At the hubstation, the composite signal is received by a dual circularly polarized antenna and the two polarizations are split into two respective banks of FM receivers. The center frequencies of the FM receivers correspond to the center frequencies of the VCOs of Figure 10. The demodulated outputs from the FM receivers reproduce the signals received at the satellite L-band antenna elements preserving their phase and amplitude relationships. Figure 11 shows an exemplary relationship between the K-band transmission spectra for the two polarizations, showing how the half-channel offset between the RHC and LHC center frequencies minimizes interaction.

Those skilled in the art will readily appreciate that the block diagrams of Figure 10 is merely illustrative of an exemplary arrangement of coherent signal transmission

according to the present invention and that many functional equivalents flow therefrom. For example, it might be advantageous first to generate the frequency modulated signals at a lower frequency than 20 GHz, for example 2-3 GHz and, after summing, to convert the composite signal to 20 GHz by mixing the summed signal with a common 18 GHz local oscillator and selecting the upper sideband with a bandpass filter.

The above discussion has centered on the coherent transportation of signals received by the satellite L-band antenna elements to the hubstation.
The same function, namely the transportation of signals generated at the hubstation, is used in reverse for radiation by respective satellite antenna elements, e.g., by the transponder of Figure 8. The hubstation can use a similar arrangement to Figure 10, but with a set of K-band frequencies different from those used in the satellite-to-hub direction, and with a larger antenna at the ground end. The satellite can employ a second, dual-polarized horn antenna for reception, or alternately use the same horn antenna and polarizer 1008 and 1009 as in Figure 10 with the addition of a transmit/receive diplexing filter for each polarization to separate the transmit and receive signals. Linear amplifier 350 can be duplicated for each polarization and used to feed respective halves of FM receiver bank 340. The same half-channel frequency offset between the carriers of the two polarizations is also advantageous in the hub-to-satellite direction.

Figure 12 shows an alternative arrangement according to another exemplary embodiment of the present invention for coherently transporting multiple signals between the hub and the satellite. In this figure, each satellite transponder channel corresponding to one antenna feed element is shown as a double downconversion process comprising an antenna filter 1200, a low-noise amplifier 1201, image rejection filter 1202, first downconvertor 1203, IF filters 1204, 1206, IF amplifier 1205 and quadrature downconvertors 1207, 1208. The first downconvertors 1203 can use the same local oscillator signal for all channels to preserve relative coherency. The quadrature downconvertors 1207 and 1208 can use the same second local oscillator cosine and sine reference signals for all channels, again to preserve relative coherency. The quadrature downconvertor outputs, for example in the 0-3 MHz range, are split in cross-over network 1209 into 0-50 KHz components on lines 1215 and 1216 and 50

KHz-3 MHz components on lines 1217 and 1218. The 50 KHz to 3 MHz components correspond to uplink traffic channels using, for example, FDMA, an FDMA-plus-narrowband-CDMA hybrid or narrowband FDMA/TDMA, and are used to modulate separate I and Q K-band transmitters for relaying these signals coherently to the hubstation. These components modulate the I and Q voltage controlled oscillators 1210 and 1211. The outputs of these oscillators are summed in a K-band summing network and the sum fed to a common TWTA for amplification to the desired downlink transmit power level. It is also advantageous to combine half the VCOS, e.g., the I VCOs into a first TWTA to form a signal for transmitting using RHC polarization, the other half being transmitted with LHC. A similar arrangement can be employed at the hubstation for coherently conveying the composite signal for each beam to the satellite.

The corresponding K-band receivers would comprise an FM receiver for each of the I signals and an FM receiver for each of the Q signals. These FM receivers would preferably have automatic frequency control (AFC) which removes DC and low frequency components of the I and Q signals, equivalent to having a notch in the frequency response in the center of the channel. This is of little consequence for wideband TDMA signals and for FDMA simply means not using the channel in the center of the band for traffic.

In the satellite, the outputs of the K-band receivers are reconstituted I and Q signals that are used to modulate COS and SIN L-band carriers using a quadrature modulator to produce coherent beam signals. These are applied to L-band power amplifiers for each beam or to the PA of the aforementioned matrix type.

The frequency arrangements used can be similar to those depicted in Figure 11, with RHC polarization being used, for example, for I components and LHC polarization for Q components, and the carrier spacings being reduced such that they are commensurate with a 3 MHz modulating signal instead of 7 MHz. A half-channel offset between the RHC- and LHC- polarized carriers is also advantageous in this I,Q method.

The I and Q signals represent, respectively, the projections of the complex received signal vectors on the real and imaginary axes, and preserving the correct amplitude relationships between the I and Q signals will then preserve the vector

relationships including relative phase. The 2n I and Q video signals can be used to frequency modulate 2n K-band carriers having less than half the channel spacing previously used in Figure 10, for example 10 MHz. While it may appear more spectrally efficient at K-band to use this method, it is difficult in practice to handle I,Q signal components down to true zero frequency due to DC offsets and frequency errors.

Consequently it is desirable to employ AC coupling and thus exclude a portion, for example 0-50 KHz, of the 0-3 MHz video signals from transmission. This places a notch in the center of the 6 MHz wide L-band bandwidth that is transponded by this exemplary method. Depending on the nature of the signals, this notch may be of no consequence. For example, in copending, commonly assigned U.S. Patent Application Serial No. _______, entitled "TDMA/FDMA/CDMA Hybrid Radio Access Methods" and filed on January 11, 1994, which is incorporated here by reference, there is disclosed a hybrid access method suitable for satellite-cellular applications in which signals are conveyed on the downlink (satellite-mobile) by wideband TDMA in which each mobile signal occupies an assigned timeslot in a repetitive frame structure, and on the uplink (mobile-to-satellite) by Frequency Division Multiple Access (FDMA) or a combination of FDMA and Code Division multiple Access (CDMA). For example, a 6.5536 megabit per second TDMA signal comprising 512 timeslots can be transmitted from the hubstation for transponding through the 6 MHz bandwidth of each satellite antenna feed element to a corresponding number of mobile phones in each cell. The omission of a small fraction of the bandwidth in the center of the channel will not disturb the character of such a signal significantly, and such disturbance as may occur can be compensated at the receiving radio using the technique for DC offset compensation disclosed in commonly assigned U.S. Patent No. 5,241,702 to Paul W. Dent entitled "D.C. Offset Compensation in a Radio Receiver" which is incorporated here by reference.

When such 512-timeslot TDMA formats are used on the downlink, one or more timeslots can be dedicated for use as a common signalling channel, also known as a calling channel, forward control channel, or paging channel. The calling channel is used by the system to broadcast calls to mobile phones originating from the network

(e.g., from a PSTN subscriber or from another mobile phone). When a mobile detects its own phone number or ID in such a broadcast message, it replies using a corresponding uplink channel commonly called the "random access channel". The random access channel is so called because it is also used by mobile phones to place mobile-originated calls, that is to request service from the network. With a large population of roaming mobile phones, these requesting events seem to the system to arise more or less at random.

According to the aforementioned "TDMA/FDMA/CDMA Hybrid Access Methods" patent application, there is associated with each downlink timeslot a corresponding uplink carrier frequency. Thus to employ the aforementioned disclosure in conjunction with the I, Q version of the present invention, the uplink carrier frequency associated with the downlink calling channel timeslots can be chosen to correspond to the +/-50 KHz in the center of the 6 MHz bandwidth and is used as the random access channel.

Accordingly, the 0-50 KHz signals from the crossover network 1209 represent random access signals and because of their relatively low bandwidth the option of digitizing on-board and transmission by digital means to the hubstation exists. This is carried out by A/D convertors 1212, the outputs from each channel of which are multiplexed in multiplexer 1213 to form a composite bitstream on the order of 60 MB/s which modulates a digital transmitter 1214 for transmission to the hubstation.

According to yet another exemplary embodiment, antenna element signals can be transported coherently between the ground station and the satellite without bandwidth expansion. Figures 13 and 14 illustrate an exemplary coherent transmission method and apparatus which is based on analog to digital conversion of each of the antenna signals followed by digital multiplexing and then modulation of the multiplexed stream on to the K-band feeder link carrier by means of Quadrature Amplitude modulation. Figure 15 illustrates an alternative apparatus derived from Figure 13 which equates to infinite AtoD and DtoA precision, thus permitting the AtoD's and DtoA's of the exemplary embodiment of Figure 13 to be replaced by analog multiplexing.

With reference to Figure 13, operation of this coherent transmission system is as follows. A 2 GHz signal received from one of a plurality of satellite-borne antenna

elements is low-noise amplified and downconverted against cosine and sine local oscillator signals using mixers 1301 and 1302. If the bandwidth at 2 GHz that is downconverted is 5 MHz, then the resulting I and Q signals are of bandwidth 2.5 MHz each. Thus the desired bandwidth of 5 MHz may be imposed by the use of 2.5 MHz cut-off low pass filters 1304, 1305 operating on the I and Q signals. These mixers, filters and AtoD convertors 1306, 1307 are repeated for each separate antenna element signal so treated. The mixers can receive the same local oscillator signals cos(wT) and sin(wT) so as not to introduce any relative phase shift between channels.

The baseband I and Q signals after filtering are converted using AtoD convertors 1306 and 1307. These are arranged to sample and convert the I and Q signals at least at the Nyquist rate, which is twice the bandwidth or, in this example, 5 MS/S. Sampling at least at the Nyquist rate allows the signals to be faithfully reconstructed from the samples. By way of example, the AtoD convertors are illustrated as having only two bits resolution, that is each I or Q signals is classified as lying nearest to one of the four values -3, -1, +1 or +3 arbitrary units, as indicated by a digital code 11, 10, 01 or 00.

In certain applications, two bits quantizing may indeed be sufficient. Such applications are characterized by the total signal-to-noise ratio in the 5 MHz bandwidth at 2 GHz being very low or even negative. This can arise, for example, when the signal bandwidth has artificially been widened by the use of coding or spread-spectrum techniques. If the signal-to-noise ratio is low or negative, a few bits resolution suffice to make the digital quantizing noise lower than the radio noise to avoid degradation. Those skilled in the art will appreciate that for applications having higher signal-to-noise ratios, more bits can be used to provide greater precision.

With the two-bit example, bit-pairs representing instantaneous I samples and Q samples are collected from all antenna elements and multiplexed using digital multiplexers 1308 and 1309. The output of digital multiplexers 1308 and 1309 is a two-bit I and two-bit Q signal, respectively, for antenna number 1, followed by the same for antenna number 2, then 3, 4, etc., until antenna 1 is again sampled. The succession of two-bit values of I and Q is then to be transmitted by modulation onto the K-band feeder link carrier frequency.

Since the number of bits per second is 4N x 5MS/S = 20N Mbits/S, a bandwidth-efficient digital modulation scheme is required to avoid the signal occupying more than the 5N MHz of the original N signals. A suitable modulation scheme can, for example, be 16 QAM. In 16 QAM, four bits of data are conveyed per transmitted symbol, by mapping two bits to one of four K-band carrier real vector values (i.e., the amplitude of a cosine carrier component) and two bits to one of four imaginary vector values (i.e., the amplitude of the sine component). The 4x4 grid of possible points that result is shown in Figure 14. Using 16 QAM, I bitpairs are mapped to the K-band I axis and Q bitpairs to the K-band Q axis using DtoA convertors 1310 and 1311. Finally, the desired K-band vector components are formed by applying the outputs of two-bit DtoA convertors 1310 and 1311 to a K-band Quadrature Modulator 1312 which is driven by K-band cosine and sine carrier waves (not shown) to form a modulated output signal for transmission via the K-band feeder link antenna (also not shown).

The multiplexer can preferably have more inputs than signals from antenna channels. For example, a typical antenna arrangement can be a hexagonal array of 61 antenna elements. A 64-input multiplexer can then be suitable, as a power of two arises naturally in multiplexer construction. The 3 spare inputs can then be connected to reference I,Q signals equal respectively to (0,0), (1,0) and (0,1). The ground station receiver can use these reference signals to synchronize its demultiplexing and to determine quadrature modulator carrier leakage (offset) from the (0,0) case, and to provide phase references from the (1,0) and (0,1) cases for discriminating the I-axis bits from the Q-axis bits.

In case two-bit quantization is inadequate, AtoD convertors 1306 and 1308 can be of a higher resolution, for example four bits. Then each 4-bit I and 4-bit Q sample will represent one of 256 possibilities, and this can be transmitted using 256 QAM in the same way as described above for 16 QAM. However, a simplification is possible by noting that the complementary operations of AtoD conversion performed in blocks 1306, 1308 and mapping to a symbol performed by DtoA convertors 1310 and 1311 simply cancel each other out and can be omitted in this alternate exemplary embodiment. Then, the full unquantized accuracy of the analog I and Q signals from

the low pass filters is preserved through the multiplexers and the digital multiplexers are replaced with analog multiplexers as shown in Figure 15.

In Figure 15, baseband signals are produced by downconvertors 1501, 1502 and low pass filters 1503, 1504 as described above with respect to Figure 13. The I,Q signals are however no longer digitized and instead are applied directly to the inputs of analog multiplexers 1505, 1506 along with corresponding signals from other antenna channels (not shown). The multiplexed I samples then modulate a K-band cosine carrier and the multiplexed Q samples modulate a K-band sine carrier, by use of quadrature modulator 1507. Spare inputs of the analog multiplexers, as previously indicated, can be used to multiplex and transmit reference values such as (0,0), (1,0) and (0,1) which can be helpful in assisting the ground station receiver to acquire demultiplexor synchronization and in correcting certain errors in the quadrature modulator such as carrier imbalance (carrier leakage, offset) and imperfect quadrature (i.e., the cosine and sine carriers are not exactly 90 degrees apart).

The configuration illustrated in Figure 15 has an advantage that substantially no bandwidth expansion of the signal takes place from 2GHz to K-band. The N, 5 MHz wide antenna signals received at 2 GHz are retransmitted at K-band using substantially the same 5N MHz bandwidth. Furthermore, no quantization noise is introduced.

A suitable analog multiplexer for the exemplary embodiment of Figure 15 can be constructed as a binary tree, in which pairs of 5MS/S signals are first multiplexed in relatively low-speed, 2-input multiplexers to form 10MS/S signals. Then pairs of these are multiplexed in higher speed 2-input multiplexers to form 20MS/S signals and so on. The multiplexers can be constructed in a bipolar, CMOS or BiCMOS integrated circuit using current steering in which a signal is applied to the junction of two transistor inputs (e.g., emitters) that are alternately enabled or disabled by a control signal (applied, e.g., to bases) to either pass the signal current through one of the devices or to shunt it away through the other. Gallium arsenide technologies, such as HBT, are also very suitable for constructing high speed multiplexers.

The ground station processing system receives the time-multiplexed antenna signals on K-band, converts those signals down to I,Q baseband signals of 2.5N MHz bandwidth each, and then demultiplexes them into N, separate 2.5 MHz bandwidth

signals of 5MS/S each (the Nyquist rate or higher). These signals can then be digitized on the ground to whatever accuracy is required for further processing such as using an equalizer for removing inter-sample interference on a sample caused by smearing from adjacent samples due to deliberate or accidental bandwidth restrictions in the K-band transmitter, receiver or propagation path. Such an equalizer operates by subtracting a defined amount of a previous and subsequent complex (I,Q) sample value from a current value, the defined amounts being given by complex coefficients that are chosen to cancel inter-sample interference. This process can also be applied in reverse for conveying to the satellite using K-band complex signal vector samples for transmission by respective antennas at, for example, S-band.

The N separate, complex (I,Q) sample streams are first preferably subjected to pre-equalizing at the ground station such that they will be received with zero intersample interference at the satellite. Then the time-multiplex modulated K-band signal is downconverted in the satellite against a K-band local oscillator to give multiplexed I and Q streams. If desired, two or more stages of downconversion can be employed so that amplification takes place at convenient intermediate frequencies. This may also apply to the 2 GHz downconverters of Figure 13, but note that the same local oscillator signals should then be employed in all corresponding stages of downconversion for each antenna element so as not to introduce relative phase shifts.

The multiplexed I,Q streams received by the satellite can be demultiplexed using the same multiplex clock (not shown) used for multiplexers 1505 and 1506. The onus is thus on the ground station to transmit a signal taking into account propagation time such that the signal will arrive in the correct timing relationship to ensure proper demultiplexing on board the satellite. In this way, the satellite function is kept simple and reliable and complexity is restricted to the ground, where equipment can be repaired should it fail.

The above description has been simplified for purposes of illustration to the case where all antenna element signals are time-multiplexed to a single TDM complex sample stream. Those skilled in the art will readily appreciate, however, that a hybrid TDM/FDM scheme could be used in which groups of time-multiplexed signals are formed and used to modulate separate FDM carriers. This modification could be used

if, for example, a single multiplex stream would result in an impractically high sample rate.

It is also for the purposes of illustration that the above description has concentrated on the Cartesian (I,Q) representation of complex signals. It is equally possible to form polar or logpolar representations of complex signals, to multiplex these signals using analog multiplexers prior to modulating a K-band feeder link or to digitize them using the method of U.S. Patent No. 5,048,059, which was earlier incorporated by reference, prior to multiplexing.

Figure 16 shows transmit signal processing in the hubstation for this exemplary embodiment of the present invention. Each voice channel to be transmitted to a mobile phone can be received either as a standard 64 KB/s PCM signal or as an analog signal which is converted to PCM. The PCM signal is then transcoded to a lower bit rate, such as 4.8 KB/s, using a conventional voice compression algorithm such as CELP (codebook excited linear prediction), RELP, VSELP or sub-band coding. The transcoded voice signal is then subject to error correction coding and supplementary bits can be added such as the Cyclic Redundancy Check bits (CRC), Slow Associated Control Channel signalling information (SACCH), per-slot syncwords and inter-slot guard symbols. This per-channel processing takes place in voice processing channel cards 1600. The output bitstreams from, for example, 500 such channel cards, are then multiplexed with a control channel data stream from a control processor (not shown) in multiplexer 1601 to form the TDMA bitstream, for example, of 6.5536 megabits per second. This is submitted to a digital modulator 1602 that numerically converts the information stream to a stream of complex numbers at a sample rate of, for example, eight samples per bit representing the I, Q components of a modulation waveform.

The TDMA signal produced as described above is targeted for transmission to a first set of, for example, 500 mobile phones in a particular cell or area. A number of other such TDMA signals formed by similar circuitry 1600, 1601, 1602 are produced for transmission to other sets of 500 mobile phones in 36 other cells. The total number of cells (e.g., 37 in this exemplary embodiment) times the number of traffic channels per cell (e.g., 500) gives the total system capacity as 18500 voice channels. The signals in timeslots 1 of each cell are transmitted simultaneously on the same frequency

to their respective cells. To avoid spillover interference from adjacent cells using the same frequency at the same time, this exemplary embodiment of the present invention includes matrix processor 1603 to process the signals from modulators 1602 by weighted addition using a matrix of 37x37 complex coefficients for each timeslot. The 37x37 coefficients for each timeslot are contained in coefficient memory 1605 which can be distributed within the components of the numerical signal processor but which is collectively identified as a separate block 1605 in Figure 16 to better illustrate the principle. During the first timeslot, a first set of coefficients C is selected from the memory and used to matrix-multiply the modulation signals from modulator 1602 to obtain signals for D/A convertors 1604. Each D/A convertor can be a dual-channel unit capable of operating with complex numbers. For example, the output signals from the matrix processor can consist each of a 12-bit real (I) and 12-bit imaginary (Q) part which are D/A converted to produce analog I, Q signals. The I, Q signals are fed to FM K-band FM transmitters for transmission from the hubstation to the satellite.

When transponded by the satellite to the ground on S-band, the result of the matrix processing will be that each mobile phone receives only its own signal, the inter-cell interference from other cells having been cancelled by the addition in the matrix processor of compensating amounts of opposite sign as determined by the coefficients retrieved from memory 1605. This is possible if the 37 mobiles using timeslot 1 in their respective cells are spatially separated, i.e., not both at the same location on the edge of their respective cells. This condition can be maintained by the exemplary timeslot assignment algorithm feature of the present invention, which also provides a general channel assignment algorithm, and is based on maximizing the signal quality provided to the worst case mobile.

A timeslot duration is typically about $40\mu S$ if a 20 mS TDMA frame period is used. One timeslot corresponds to 256 bit periods at 6.5536 MB/s and 2048 complex numbers are produced by modulator 1602 for every timeslot. After matrix processor 1603 has processed 2048 sets of 37 complex number inputs using the set of coefficients for the first timeslot, the coefficients are changed for the second and for subsequent timeslots to effect correct interference cancellation between corresponding sets of 37 mobiles using timeslots 2, 3 etc.

If two mobiles receiving the same timeslot in different cells approach each other too closely during the progress of a conversation, the control processor (not shown) will note a difficulty in arriving at a suitable set of coefficients for interference cancellation. This is highly unlikely given the limited speed of landmobile phones in relation to typical cell size, but if it occurs, the control processor evaluates whether a timeslot change would be appropriate for one of the mobiles. The aim is to connect the mobile using a timeslot that no other mobile in close proximity is using. If necessary, a mobile even occupying an ideal (e.g., low interference) timeslot could be shifted to a just adequate (e.g., barely tolerable interference) timeslot to release its original timeslot to solve a proximity problem at hand. It is probably not necessary in practice to consider such a situation because with, for example, 500 timeslots to choose from, it would usually be possible to find a better timeslot than the timeslot currently threatening to cause bad signal quality. Allowing one timeslot change per cell per 10 seconds, for example, would be expected to achieve adequately rapid optimization of timeslot assignments and adequate adaptation to mobile movement.

In fact a more rapid rate of adaptation is provided to handle the rate at which new calls are placed and old calls cleared down. With a capacity of 37 mobiles per timeslot and an average call duration of 3 minutes, a particular timeslot is vacated in some cell approximately every 5 seconds and a new call is then assigned to that timeslot. Overall, given, in this example, 500 timeslots and 37 cells, 100 timeslots spread over all the cells are vacated every second and reassigned.

Such a communications system should be designed to not be loaded up to 100% of system capacity or the next call attempt will be blocked. With 500 timeslots per cell available, an average loading of 474 timeslots can be reached for a blocking probability of 1%. Thus, on average, 26 out of 500 timeslots are unused on each of the 37 multiplexers 1302 in this exemplary embodiment. It should be noted that it is immaterial which multiplexer is used to transmit a particular timeslot to a mobile. Whichever timeslot is selected, it is the choice of an associated column of matrix coefficients that determines that mobiles using the same timeslot are non-interfering. Thus if the same timeslot, for example number 371, is vacant on two or more multiplexers 1601, it is immaterial which one is used to connect a new call.

Thus the assignment algorithm executed in the control processor first determines which timeslot is vacant on the greatest number of multiplexers. This is the timeslot on which there are currently the least number of mobile conversations. Using information from the random access receiver on the relationships between signals received from the new mobile (i.e., the C-matrix coefficients determined by correlation of the new mobile's random access signal with all antenna element signals), the control processor evaluates the change needed to the set of coefficients in coefficient memory 1605 associated with the vacant timeslot to maintain non-interference if that timeslot were to be used for the new signal.

The general principles that explain how the choice of coefficients in coefficient memory 1605 is arrived at for an exemplary embodiment will now be outlined.

As discussed earlier, for receiving signals from mobiles, antenna element 1 received an amount C11 of mobile signal Ml plus an amount C12 of mobile signal M2 and so on. To state this more generally, antenna element k received an amount Cki of mobile i's signal. Assuming reciprocity, a signal Tk transmitted from antenna element k would be received in an amount Cki.Tk at mobile i, because the path from element k to mobile i is assumed to have the same attenuation and phase shift in each direction, given by the complex number Cki.

Therefore the signals $\underline{R}$ received at the mobiles are related to the signals transmitted by the antenna elements by the matrix equations:

$\underline{R} = C^t.\underline{T}$; where the superscript t indicates a transpose matrix.

The transpose of C is used because the first index k of Cki multiplies the corresponding index of the T-element, while, in the mobile-to-satellite direction where the signals received at element k from mobile i are given by Cki.Mi, it is the second index i of C which corresponds to the index i of the mobile signal Mi that it multiplies. Thus the indices of the matrix coefficients are transposed in the satellite-to-mobile direction as compared to the mobile-to-satellite.

In order to achieve non-interference, the set of signals transmitted from the satellite antenna elements should be given by:

$$\underline{T} = C^{t-1} \cdot \underline{R}$$

The inverse of the transpose is just the transpose of the inverse, therefore the set of coefficients contained in coefficient memory 1605 for downlink timeslot(j) are just the transpose of the set of coefficients associated with uplink frequency(j) in numerical processor 650 of Figure 9, at least under the assumption of reciprocity.

Reciprocity applies when the uplink and downlink frequencies are the same. Relative amplitude reciprocity applies if the antenna element patterns are the same on both uplink and downlink frequencies. Phase reciprocity does not apply, because relative phase depends on the small differences in relative distance travelled by the signals to/from each element, divided by the wavelength and multiplied by 360 degrees. If the wavelength is different on the uplinks and downlinks, then the phase relationships will be different. However, relative time delay differences are frequency independent and therefore reciprocal. Accordingly, a set of relative phase differences at one frequency can be translated to a set of time differences using a first wavelength, and then reconverted to a set of phase differences using another wavelength in order to derive a set of coefficients valid at a second frequency from a set known at a first frequency.

Based on the foregoing discussion, the coefficients for transmit contained in memory 1605 according to an exemplary embodiment of the present invention can be determined by the following steps:

(1) correlating the signal received from a new mobile during its random access transmission with the individual antenna beam element signals to determine a new column of coefficients for the receive C-matrix;

(2) determining a new inverse C-matrix for receiving traffic from the new mobile based on the old inverse C-matrix and the new column;

(3) transforming the new receive C-matrix column to a new transmit C-matrix row by scaling relative coefficient phase angles using the ratio of up- to down-link frequencies; and

(4) determining a new transmit inverse C-matrix based on the old transmit inverse C-matrix and the new row.

An exemplary detailed mathematical procedure which can be used to carry out the above exemplary embodiment is now developed for the underloaded case, i.e., the case when there are fewer currently active mobile signals than the number of antenna feed elements available on the satellite to communicate with them. Such spare capacity is typically designed for in radio telecommunications to provide a 98% probability of having a free channel to serve a new call so that customers are not overly irritated at call blockages.

The active mobiles are designated 1.....m and the signals intended to be received by these mobiles are designated R1.....Rm for this example. The antenna element/transponder channels available for communications therebetween are designated 1....n and the signals fed to the antenna elements for transmission by each respective element are designated T1.....Tn. As before, the matrix C, this time an m x n non-square matrix, determines how much of each transmitted signal Tk reaches each mobile as Ri, the matrix is given by the equations:

$$
\begin{aligned}
R1 &= C11.T1 + C12.T2...... + C1n.Tn \\
R2 &= C21.T1 + C22.T2...... + C2n.Tn \\
\cdot \quad &\quad \cdot \qquad \quad \cdot \qquad \qquad \quad \cdot \\
\underline{R}m &= Cm1.T1 + Cm2.T2...... + Cmn.Tn
\end{aligned}
$$

or simply $\underline{R} = C.\underline{T}$ in matrix/vector notation.

Because C is no longer square, it has no direct inverse, so there is no unique solution for T given by:

$$T = C^{-1} . R$$

Instead there are a continuum of solutions, as we have more degrees of freedom to choose T values than conditions to satisfy (i.e., $n > m$).

However by imposing the condition that the mean square power fed to the antenna elements in order to create the desired mobile received signals R shall be minimized, the unique solution below is obtained:

$$T = C^t . (CC^t)^{-1}$$

This equation can be derived as follows. Let $R_{desired}$ be the M-element vector of signals we wish to be received at the receiving stations, and T be the N-element vector of signals applied to the transmitting antennas, where N > M. C is an M by N matrix of coefficients Cik that describes how the signal from transmitter antenna j propagates to receiving station i. Denoting by $R_{achieved}$ the M-element vector of signals actually received, we thus have

$$R_{achieved} = C.T \text{ ....(1)}$$

We wish to find what T should be as a linear function of the signals desired to be received, so that the minimum total transmit power is consumed in the process. The linear combinations formed by the coefficients of an M by N matrix A to be found are :

$$T = A.R_{desired} \text{ ....(2)}$$

Substituting for T from (2) into (1) we get:

$$R_{achieved} = C.A.R_{desired}$$

showing that $R_{achieved} = R_{desired}$ only if C.A is the MxM unit matrix I

$$\text{Thus } C.A = I \text{ ....(3)}$$

is a necessary condition. Since C is not square, we cannot simply invert it and write:

$$A = C^{-1}$$

Moreover, C.A. = I is a set of MxM equations that the NxM unknowns A must fulfill so that the MxM terms of the product indeed give the MxM unit matrix I.

Since N > M, the number of unknowns is greater than the number of equations, so there is no unique solution to equation (3), but a continuum of solutions. Other conditions must be imposed to define a particular solution of interest. The condition imposed here is that the total power inherent in transmitting the vector of signals T is minimized.

It can be verified that a particular solution of equation (3) is $A = C'(CC')^{-1} = U$, where ' signifies conjugate transpose. This can be verified by substituting the particular solution U for A in equation (3), obtaining:

$$C.U = C.C'(CC')^{-1} = (CC')\cdot(CC')^{-1}$$

,which is clearly equal to I as required.

A general solution can be formed by adding an arbitrary matrix V to the particular solution found above, obtaining:

$A = C'(CC')^{-1} + V$, but this must still fulfill equation (3). Substituting this value of A into equation (3) we get:

$C(C'(CC')^{-1} + V) = I$

i.e., $CC'(CC') + CV = I$

i.e., $I \qquad + CV = I$

i.e., $CV = 0$ ....(4)

Thus, V may be arbitrary only so long as it fulfills equation (4). It is possible for a non-zero V matrix to give identically zero when premultiplied by C as long as all V's columns are orthogonal to all C's rows. The rows of C are N-element vectors, but there are only M of them, therefore they do not totally span their N-dimensional space. There are N-M other dimensions in that space that the rows of C do not project into, and the columns of V may thus consist of any vectors that are confined to that N-M dimensional sub-space that do not project into C's M-dimensional sub-space.

Thus, the general solution of equation (3) is

$A = U + V$; where U is the particular solution identified above and V must satisfy $C.V = 0$

The transmitted signals T are given by

$T1 = A11.R1 + A12.R2 + \dots\dots\dots\dots A1m.Rm$
$T2 = A21.R1 + A22.R2 + \dots\dots\dots\dots A2m.Rm$

. . .

. . .

$Tn = An1.R1 + An2.R2 + \dots\dots\dots\dots Anm.Rm$

where R1, R2, etc. are the elements of $R_{desired}$.

If R1, R2, etc. are all independent signals intended for different receiving stations, there is no correlation between them so they add rms-wise in the linear summing process that forms the T-elements.

Thus, the mean square value of T1 is just $|A11.R1|^2 + |A12.R2|^2 \dots + |A1m.Rm|^2$

Likewise, the mean square value of T2 is $|A21.R1|^2 + |A22.R2|^2 .... + |A2m.Rm|^2$

Adding these expressions down columns that contain the same Ri, we get:

POWER = $SIGMA_j \left[|Ri|^2 \cdot SIGMA_i|Aij|^2\right]$

Now $SIGMA^i|Aij|^2 = SIGMA_i(Aij.Aij) = SIGMA_i\,(A'ji.Aij)$, where A'ji refers to element ji in the conjugate transpose of A.

But this value SIGMA is simply the jj diagonal term of the whole matrix product $Xjk = SIGMA_i\,(A'ji.Aik)$, which is the equation for matrix multiplying A' and A, i.e., X = A'A.

Now substitute A = U + V; then:

X = (U'+V').(U+V) = U'U+V'V+U'V+V'U and

U'V+V'U = 2 Re (U'V).

Substituting $U = C'(CC')^{-1}$, i.e., $U' = (CC')^{-1}.C$, into the foregoing equation, we get $U'V = (CC')^{-1}.CV = 0$ because CV=O.

Therefore 2Re(U'V) = 0 and U'V+V'U = 0.

Hence $SIGMA_i|Aij|^2 = SIGMA_i\,(|Uij|^2 + |Vij|^2)$, leading to:

POWER = $SIGMA_j \left[|Ri|^2 . SIGMA_i|Uij|^2\right] + SIGMA_j \left[|Ri|^2 . SIGMA_i|Vij|^2\right]$

Since the two terms involving respectively U and V can only be positive, the power is minimized when the choice of the arbitrary matrix V in the second term is zero. Hence, the solution for the transmit signals that create the desired received signals is:

$T = A.R_{desired}$ where $A = C'(CC')^{-1}$

This solution also holds for the case where N=M, for then C is square and the above reduces to:

$A = C^{-1}$

Applying the foregoing principles, the spare degrees of freedom are used not just to create co-channel interference free signals at every mobile, but also to maximize the wanted signal values for a given total radiated power. The

total mean square radiated power is in fact the sum of the square magnitudes of the coefficients of the matrix A defined by:

$$A = C.^{t}\ (C.C)^{-1}$$

The sum of the squares down a column of A gives the radiated power used in communicating with a corresponding mobile. The worst case mobile, i.e., that using the most satellite power, can thus be identified. According to an aspect of the present invention, the control processor at the hubstation periodically examines whether total satellite power can be minimized (or utilization of power optimized) by removing the worst case mobile from the current group the mobile is associated with and associating that mobile with a different group. This is done by recomputing the above expressions with C diminished by the row corresponding to the worst case mobile, thus determining the satellite power saving that would be saved in supporting only the remainder in the most efficient manner. Then the removed row of C is used to augment in turn each of the C matrices associated with other groups of mobiles using different frequency channels (FDMA) or multi-carrier (CDMA) or timeslots (TDMA) and the above expressions computed to determine the increase in power that would be necessary to support that mobile as a member of each of the other groups in turn. If the increase in power in one of these cases is less than the power saved by removing the mobile from its original group, then a frequency or timeslot handover to the new group can be performed in order to improve satellite power utilization. This procedure can likewise be used for determining which of a number of existing groups a new mobile call should be associated with, i.e., to find the group that would result in minimum increase of satellite power used when a new call is connected.

Figure 17 shows an exemplary embodiment illustrating the interconnections between the transmit and receive matrix processors and the control processor at the ground station to effect the above-described interference cancelling and optimum channel allocation behavior.

The receive matrix processor 1700 receives digitized signal samples from the ground station RF section. The receive processing can be structured, for example, according to the exemplary FDMA embodiment of the invention of Figure 9, or according to the exemplary TDMA embodiment of Figure 16. Moreover, an exemplary CDMA embodiment can be constructed by, for example, increasing the bandwidth of the channel splitting filters and including a CDMA version of the per-channel processing in the circuitry of Figure 9. Further, exemplary embodiments of the present invention may be constructed using the novel subtractive CDMA system described in U.S. Patent 5,151,919 to Paul W. Dent entitled "CDMA Subtractive Demodulation", which is incorporated here by reference. These features of the present invention also lend themselves to implementation in land-based cellular systems.

The receive matrix processor 1700 separates the individual channel signals by applying inverse C-matrix coefficients supplied by control processor 1702 as described above, to eliminate or suppress co-channel interference. These coefficients can, for example, be determined as follows.

When M spatially separated antenna/receiver channels receive different combinations Ri of M signals Si, given by

$Ri = SIGMA_j (Cij.Sj)$ ....: (5)

or in matrix notation, $R = C.S$, then the separation of the M signals has a straightforward solution $R = C.S^{-1}$ .... (6)

When the number of antenna/receiver channels N is greater than the number of signals M they receive, the matrix C is not square and cannot be inverted. There are a continuum of solutions possible using any subset M of the N channels, but there can also be a desired unique solution.

The reciprocal problem for transmitting M signals using N transmitter channels was solved above by imposing the additional desire to minimize total transmit power. In the receiving case, we can find the desired unique solution by imposing the condition of maximizing the signal to noise ratio.

To do this, a finite amount of noise must be assumed to exist in the receivers.

Before this solution is described, another solution will be described for solving the equations:

$$\begin{array}{l} C11.S1+C12.S2 \ldots\ldots\ldots\ldots +C1M.SM = R1 \\ C21.S1+C22.S2 \ldots\ldots\ldots\ldots +C2M.SM = R2 \qquad \ldots\ldots\ldots\ldots(7) \\ \quad . \qquad . \quad \ldots\ldots\ldots\ldots \quad . \quad . \; . \\ CN1.S1+CN2.S2 \ldots\ldots\ldots\ldots +CNM.SM = RN \end{array}$$

When N > M there is an excess of equations over unknowns. They should all be consistent and solving any subset M of N should yield the same answer. Due to receiver noise, however, which causes uncorrelated errors in the received values R, the equations will not all be exactly consistent.

A known solution to this is the so-called least-squares solution. The least squares method seeks the solution which minimizes the RMS sum of the noise errors needed to be added to the R-values to make the equations consistent.

An error vector E may be defined as E=C.S-R .... (8)

The sum square error is then

$$E'E = (C.S-R)'.(C.S-R) = S'.C'.C.S - R'.C.S - S'.C'.R + R'.R \;\ldots\ldots\; (9)$$

Differentiating this expression with respect to each R value to obtain the gradient yields:

$$grad(E'E) = 2C'.C.S - 2C'.R \;\ldots\ldots\ldots\ldots\ldots\ldots\ldots\; (10)$$

E'E is a global maximum where grad(E'E) = 0,

$$\text{i.e, } C'C.S = C'R, \text{ or } S = (C'C)^{-1}.C'.R \;\ldots\ldots\ldots\ldots\ldots\; (11)$$

The least squares solution for the M signals is thus S = A.R where

$$A = (C'C)^{-1}.C' \;\ldots\ldots\ldots\ldots\ldots\; (12)$$

This may be compared with the least-power transmit solution where

$$A = C'(CC')^{-1}$$

The least squares solution for reception given above is not necessarily that which maximizes the quality of each signal. To find the solution that maximizes each signal quality we in turn find the best A matrix row that yields that signal.

Separated signal $\underline{S}i$ is given by row i of A, henceforth written Ai, multiplied by the vector of receive channel outputs R, i.e., $\underline{S}i$ = Ai.R. R is given by C.S + Noise where "Noise" is a vector of uncorrelated noise having components N1,N2 .... in the receiver channels.

Thus $\underline{S}i$ = Ai.C.S + Ai.Noise .... (13)

The amount of wanted component Si that appears in $\underline{S}i$ is given by

(Ai1.C1i + Ai2.C2i + Ai3.C3i ...... + AiN.CNi).Si = Ai.Ci where Ci means the ith column of C.

Assuming all Si are transmitted with unit power, the power in the extracted wanted component is

$P = |Ai.Ci|^2 = Ai.Ci.Ci'.Ai'$ .... (14)

There are also, however, unwanted components in the extracted signal due to the other signals Sk. The sum of the unwanted powers for all k not equal to i is given by

I = Ai.Cdim.C'dim.Ai' .... (15)

where Cdim means the matrix C with column i removed.

In addition, there is a noise power given by

$|Ai1.N1|^2 + |Ai2.N2|^2 ...... = AiAi'.nI$ .... (16)

where n is the mean square value of each of the noise signals N1,N2, etc.

The signal-to-noise-plus interference ratio is then given by

$$\frac{P}{I+N} = \frac{A(Ci.Ci')A'}{A(nI+Cdim.C'dim)A'} \qquad \text{..........................} \; (17)$$

Mathematicians will recognize this expression as the ratio of Hermitian forms. The maxima and minima of such expressions as

$$\frac{X'UX}{X'VX}$$

are given by the eigenvalues q of $V.^{-1}U$, i.e., by the solution of

$\det(V^{-1}.U-qI) = 0$. The values of X which give these extrema are the corresponding eigenvectors. $V.^{-1}U$ in our case is $(Cdim.C'dim+nI)^{-1}.Ci.Ci'$ and X is A'.

We now use the theorem that the eigenvalues of the product of an n by m matrix with an m by n matrix where $n>m$ are equal to the eigenvalues of the product taken in reverse order, plus n-m zero eigenvalues.

Using as the two matrices in question the N by 1 matrix $(Cdim.Cdim'+nI)^{-1}.Ci$ on the one hand and the 1 by N matrix Ci' on the other hand, the eigenvalues we need must be those of the inverse product

$$\underset{1xN}{Ci'}.\underset{NxN}{(Cdim.C'dim+nI)^{-1}}.\underset{Nx1}{Ci} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (18)$$

This, however, has dimension 1x1, i.e., it is a scalar, so it has only one non-zero eigenvalue.

Hence $q = Ci'.(Cdim.C'dim+nI)^{-1}.Ci \quad \ldots\ldots\ldots\ldots\ldots (19)$

The associated eigenvector Ai' is the solution V of an equation of the form

$$Matrix.V = V.Eigenvalue$$

$$(Cdim.C'dim+nI)^{-1}.Ci.Ci'.V = Vq \quad \ldots\ldots (20)$$

Substituting for q from equation (19),

$$(Cdim.C'dim+nI)^{-1}.Ci.Ci'.V = V.Ci'.(Cdim.C'dim+nI)^{-1}.Ci \quad \ldots\ldots\ldots (21)$$

It may be verified that letting $V = (Cdim.C'dim+nI)^{-1}.Ci$ makes the right hand and left hand sides of equation (21) identical. Thus, this eigenvector is the optimum solution for the row of coefficients Ai that extract Si from R with best signal-to-noise+interference ratio.

If instead we set out to maximize signal to (signal+noise+interference) ratio we would get

$$Ai' = (C.C'+nI)^{-1}.Ci \text{ or } Ai = Ci'.(C.C'+nI)^{-1} \quad \ldots\ldots\ldots\ldots\ldots (22)$$

i.e., the whole C-matrix is used in the inversion and not Cdim with one column removed. The value that maximizes S(S+N+I) should, however, be the same as that which maximizes (S/(N+I) as their reciprocals differ by the constant 1 only.

It can be shown that this solution only differs by a scalar factor $1/(1+q)$ from the solution which maximizes $S/(N+I)$, and since a fixed scaling does not change signal to noise ratios, it is effectively the same solution. If such Ai's are now derived for all i and laid one under one another to form an M by N matrix A, the rows Ci', being the original columns Ci, also lie under one another to form the matrix C'.

$$\text{Thus } A = C'(CC'+nI)^{-1}$$

This is similar to the solution for minimum transmit power derived above, except that the "C" matrix here is the transpose of the transmit matrix and is M by N instead of N by M. That means that NxN matrix CC' has rank of only $M < N$ and it has no direct inverse, being singular. However, the addition of the noise down the diagonal through the term nI is the catalyst that makes the matrix to be inverted non-singular and the above solution computable.

The solution in the transmit case provided a way to test how much the total transmit power, having been reoptimized, would have to increase to support one extra signal. Reciprocally, in the receive case, it can be tested how the addition of a new signal to those already received would affect the signal to noise ratio after re-optimization of the above coefficients with an extra column added to the C matrix. The extra column of C's in question represents the relative strengths and phases which the new signal is received by the N receiver/antenna channels. This is determined while the new signal is appearing on the random access channel and not in conflict with other signals. Furthermore, random access can be made with higher power or more coding than for normal traffic so as to facilitate detection and decoding.

The signal is decoded and retrospectively the decoded signal can be correlated with signal samples recorded from each of the N channels to determine the new C-matrix coefficients. A test is then made by appending the new C column to each of a number of candidate C matrices in turn associated with different groups of ongoing signals in order to determine the

group that would have its worst case SNR degraded the least by inclusion of the new signal. This then determines the allocation of a channel to the new signal for traffic, and explains how the C matrix coefficients are arrived at a row-at-a-time during the random access and channel allocation process.

The separated channel signals are processed in separate channel processors 1701. The channel processors can either be engaged to process subscriber traffic, after a call has been connected, or can be employed to search for random access signals from a given direction. The latter is done by combining the received signals from the satellite to form beams covering fixed regions of the earth from which random access signals may be received. The coefficients used may be chosen by the control processor 1702 to provide cancellation or reduction of interference from other signals on the same frequency from other regions so as to maximize the probability of intercepting a random access message. The random access message can also be provided with an additional degree of error correction coding to maximize reception probability in the absence of a-priori knowledge of the direction from which an access attempt is received. Optionally, the random access channel can be frequency-planned to avoid immediate frequency re-use in adjacent cells, for example by the use of a 3-cell frequency or timeslot re-use plan, since using three frequencies or timeslots for random access does not have such a deleterious effect on total system capacity as if such a frequency usage plan were adopted for every traffic channel.

The channel processors 1701 provide information to the control processor 1702 regarding the amount of each signal in each beam channel or separated channel, which the control processor 1702 uses to control the interference cancellation coefficients used by the receive matrix processor 1700. Depending on, for example, the determination respectively of correlations between each separated signal and each beam signal or the determination of correlations between separated signals, two different control concepts can be applied by control processor 1702.

In a first exemplary control implementation, a separated channel signal decoded by a channel processor 1701 is correlated or partially correlated with each non-separated beam signal in turn. The electrical connections for achieving this correlation are disposed between every channel processor 1701 and every other channel processor 1701, however these connections are omitted from Figure 17 for clarity. The part of the separated signal that is used for correlation can suitably be a known bit pattern in the channel signals, for example a synchronization word or bit pattern. The correlation results directly represent the C-matrix coefficients and these are processed by the control processor to obtain A-matrix coefficients as defined above.

In a second exemplary control implementation, a separated channel signal decoded by a channel processor 1701 is correlated with at least part of other channel signals to determine the residual amount of non-cancelled interference present due to other channel signals. That part of the other channel signals with which correlation is performed can suitably be a known pattern contained in each signal, such as a synchronization word. Since these patterns are known it is not necessary to cross-couple the channel processors 1701 to each other, thus avoiding a mass of interconnections. Furthermore, since adaptation of the receive matrix coefficients by the control processor 1702 does not have to take place at a rapid rate, as they are relatively static for a given set of transmitting mobile phones, the correlation with different signals can occur at different times at which the transmitters, by prearrangement, insert a special sync word for the purpose of correlation.

For example, suppose a known, 16-bit sync pattern is employed within each segment of transmitted signal, e.g., a TDMA timeslot. There are 16 possible orthogonal 16-bit words, so 16 different signals can be allocated orthogonal sync words. A Fast Walsh Transformer such as the one described in U.S Patent Application No. 07/735,805 entitled "Fast Walsh Transform Processor" and filed on July 25, 1991, which is incorporated here

by reference, provides an efficient means to correlate a signal simultaneously with all possible orthogonal codewords and thus directly determine the residual, non-cancelled interference amounts. If however the number of signals whose residual interference contributions are to be discriminated is greater than 16, for example 37, then 15 at a time can be arranged to use different orthogonal codewords while the other 22 use the 16th codeword. The 15 which are chosen to use different codewords can be changed between successive TDMA frames such that after slightly over two frames all signals have been uniquely discriminated.

This exemplary procedure can also be applied to FDMA or CDMA uplink modulations. In the CDMA case, for example, orthogonal spreading codes can be allocated to facilitate discrimination. If a hybrid FDMA/CDMA uplink is used with, for example, four overlapping, orthogonal CDMA signals on each frequency channel as described in the aforementioned disclosure, U.S. Patent Application Serial No. _______, entitled "TDMA/FDMA/CDMA Hybrid Radio Access Methods" and filed on January 11, 1994, then the system can readily search simultaneously for known sync patterns employing all four orthogonal codes. By permuting the underlying sync patterns as described above, it is possible to discriminate residual interference contributions from any number of different CDMA transmissions using the same channel frequencies at different locations. This can be accomplished, for example, after separating the signals using the C-matrix, a signal can be correlated with its own known bit pattern and the known patterns of other signals that should have been cancelled; results of the latter correlations yield the amount of residual, uncancelled signal and can be used to update the C-matrix.

In this second exemplary implementation, C-matrix coefficients are not directly determined, but rather the residual interference amounts are related to errors in the A- and C-matrix coefficients. This relationship can be demonstrated as follows.

The satellite or base station broadcasts N combinations of M desired signals from N transmitter/antennas. The N combinations should be chosen such that each of the receiving stations receives only its intended signal, and the other M-1 at that receiver are cancelled. The N linear combinations are preferably those derived as set forth above, which result in each receiving station receiving its intended signal only, and with minimum total transmitter power.

The transmitted signals $T = \begin{bmatrix} T1 \\ T2 \\ . \\ . \\ TN \end{bmatrix}$ are formed from the signals desired to be received $R_d = \begin{bmatrix} Rd1 \\ Rd2 \\ . \\ . \\ RdM \end{bmatrix}$ by multiplying the vector Rd by the N by M matrix A, i.e., T = A.Rd.

A is in turn shown above to be preferably equal to $C'(CC')^{-1}$ where Cij is the propagation from transmitter/antenna j to receiver i. Estimates of Cij are made at call set-up time for the receive direction and transformed to estimates for the transmit direction as described above. There will, however, be errors in the estimates of the Cij for the transmit direction that are used to compute the matrix A. Let us assume that the estimated transmit matrix C is equal to the true matrix Co plus an error matrix dC, i.e.,

$$C = Co + dC \text{ or } Co = C - dC$$

The signals Ra actually received by the receiving stations are given by the true C-matrix Co times the transmitted signals, i.e.,

$$Ra = Co.T = Co.A.Rd = (C\text{-}dC)C'(CC')^{-1}.Rd = Rd - dcC.C'(CC')^{-1}.Rd$$
$$= Rd - dC.A.Rd$$

The errors dR in the received signals dR = Rd-Ra are thus given by

$$dR = dC.A.Rd \quad ...... (23)$$

Each error element i of the error vector dR contains a part e_{ij} of each of the other unintended signals j.

If the M signals contain known signals, patterns or syncwords, by correlating with these at a mobile receiving signal i, it is possible to determine the residual unwanted amount of signal j, and thus determine e_{ij}.

The syncwords can be orthogonal so that correlation with all of them can be performed at the same time by means of an orthogonal transform such as the Walsh-Hadamard transform. If the number of orthogonal codewords available is less than the number of signals M, the orthogonal codewords can be assigned to groups of immediately surrounding beams or cells whose signals are most likely to interfere due to imperfect cancellation. A limit set of orthogonal codewords can be permuted between the M signals to allow different subsets to be resolved at a time, and all M to be resolved sequentially. In this way, by correlating the received signals Ra over the portion containing the known signal pattern with all orthogonal codewords the amount of own codeword is obtained as well as the amount of unwanted codewords. The amount of other codewords is scaled by dividing by the complex amount of own codeword correlation to yield the normalized error residuals e_{ij} that may then be complex-value averaged over several measurement intervals before being reported by the receiving stations back to the transmitting stations on a reverse Slow Associated Control Channel. To reduce the volume of reporting, each mobile can at each interval restrict itself to reporting only the largest error its correlator determines. The transmitting station can optionally either assume that the other errors are

zero at that station, or that they are as previously reported if no action to correct them has been taken in the meantime.

The matrix $E = e_{ij}$ may thus be equated to the matrix dC.A in (23), so we have $dC.A = E$ or $A'.dC' = E'$.

This is an insufficient set of equations for the unknowns dC', but a unique solution exists for which the sum of the squares of the dC's is least, that solution being

$$dC' = A(A'A)^{-1}.E'$$

Moreover, if $A = C'(CC')^{-1}$ then $A(A'A)^{-1} = C'$, therefore $dC' = C'.E'$ or

$$dC = E.C$$

Thus, given the original estimate of C and the residual correlation measurements reported by receiving stations, the error dC in the original estimate may be calculated and the estimate of C gradually refined.

As mentioned above, if the reverse SACCH signalling capacity does not allow all errors to be reported every time, it is sufficient to report only the largest. The transmitter can choose only to correct the largest there and then, or to wait until others are reported. In order to ensure that others are reported, the transmitter can request the receiver to make specific measurements via the forward SACCH channel. These refinements are mentioned for the sake of completeness in describing the scope of the invention, but the extra complexity is probably not needed in a satellite-mobile communications system where the relative positions of mobiles in the satellite beams changes only slowly relative to the speed of communications.

The control processor obtains initial estimates of the downlink C and A matrix coefficients, measured on the uplink by syncword correlation as described earlier, to the downlink frequency. The control processor then continually outputs corrected A-matrix coefficients suitably translated to the downlink frequency as described above to transmit matrix processor 1704.

A complication can arise in performing this translation due to phase mismatches between each antenna element channel. It was stated above that the relative amplitude between signals on the uplink and downlink